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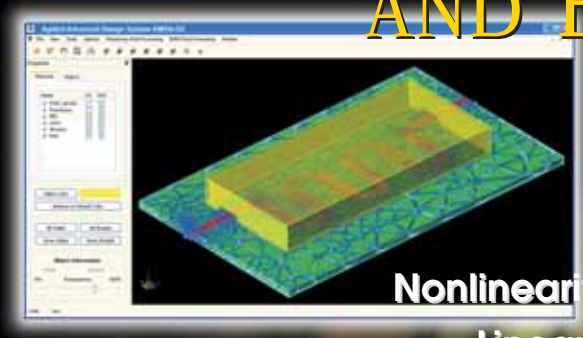

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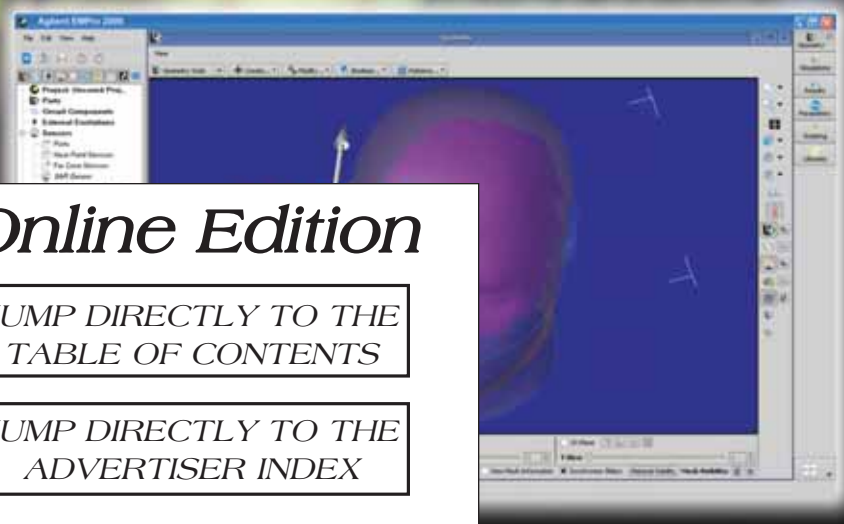
Featured Products—Transistors, Wireless Test, Military & Space



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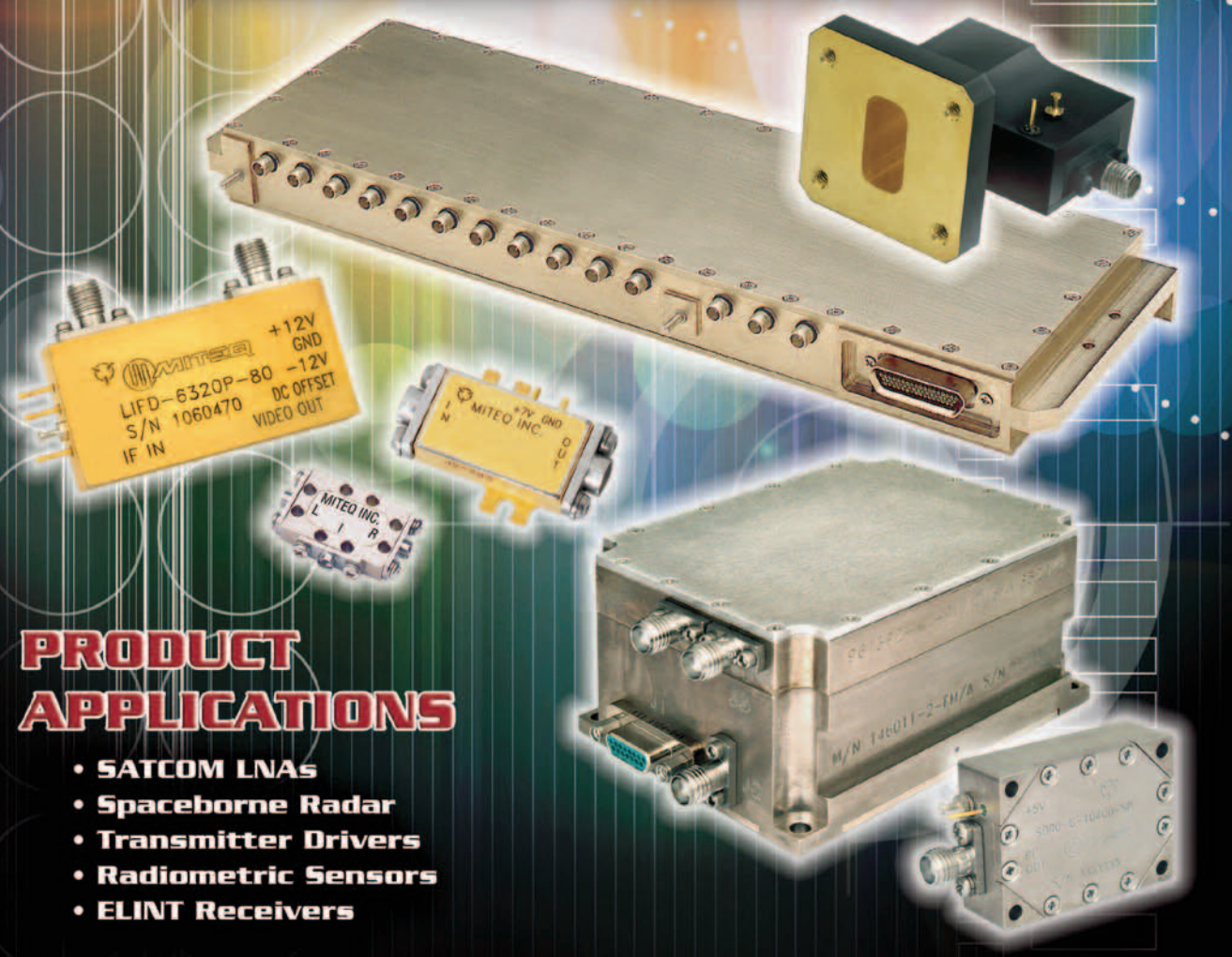

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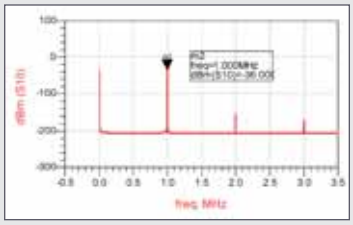

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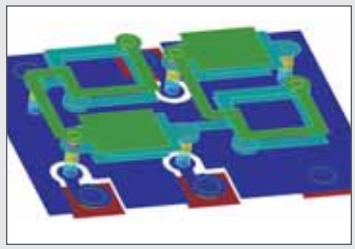

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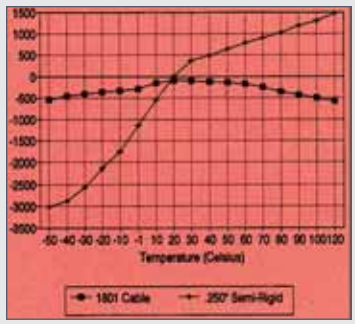

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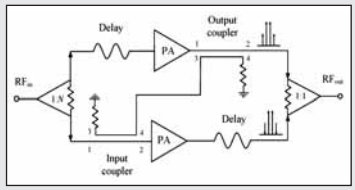

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On the cover—On page 38, readers can find the story behind EMPro, Agilent's new 3-D electromagnetic simulation tool that integrates time- and frequency-domain EM analysis capabilities with the company's Advanced Design System (ADS). (*Artwork provided by Agilent Technologies*)

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Editorial Director
Gary Breed
gary@highfrequencyelectronics.com
Tel: 608-437-9800
Fax: 608-437-9801

Publisher
Scott Spencer
scott@highfrequencyelectronics.com
Tel: 603-472-8261
Fax: 603-471-0716

Associate Publisher
Tim Burkhard
tim@highfrequencyelectronics.com
Tel: 707-544-9977
Fax: 707-544-9375

Associate Editor
Katie Landmark
katie@highfrequencyelectronics.com
Tel: 608-437-9800
Fax: 608-437-9801

Business Office
High Frequency Electronics
7 Colby Court, Suite 7-436
Bedford, NH 03110

Editorial and Production Office
High Frequency Electronics
104 S. Grove Street
Mount Horeb, WI 53572

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Subscriptions
Sue Ackerman
Tel: 651-292-0629
Fax: 651-292-1517
circulation@highfrequencyelectronics.com

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3G/4G Technology: Does the Standard Really Matter?

Gary Breed
Editorial Director

Behind all the marketing and advertising of "mobile broadband" lie several different transmission technologies, including LTE, EDGE, W-CDMA, TD-SCDMA, HSDPA, WiMAX, some variations on these basic schemes, plus new systems still in development. Each method has its supporters, and each has performance differences compared to the others. The question that often comes up is, "Which one is best?"

My first guess at an answer is, "None of them." First, they all deliver much higher data speeds than the technologies they replace. Next, there are enough differences among them that any serious observer would see that each has its strengths and weaknesses. Some provide higher data rates; some offer a more straightforward evolution from systems already in operation.

My next guess at an answer is, "All of them." Of course, proponents of each technology can point to certain specs and claim that theirs is the best. (And boy, do they make those claims loudly!) That's not really my concern—what I mean by "all of them" is that demand for higher speed mobile data services is growing, and wireless system operators need to keep working to get the technology deployed, regardless of which standard they have chosen for implementation.

There is some concern that having different systems available is a problem, but here in the U.S., where the distribution of handsets is mainly done by the operators and their agents, compatibility is a minor issue. Other parts of the world have uniform standards for their original voice-based systems, but even some of these countries/regions are allowing service providers to use different technologies for broadband communications. Even when a single standard is mandated, the chosen standards are different from one wireless market to another.

Fortunately, the issue of multiple standards (plus variants) is also being addressed by the handset and mobile appliance makers—with software-adaptable radio and baseband technologies that allow one handset, PDA or other device to handle multiple standards. These flexible devices may even allow updates and reconfiguration of networks in the future, making the choice of transmission technology almost irrelevant. Weaknesses can be addressed by changing the modulation or data format "on the fly" without the user even knowing about it. Such a universal

phone would also reduce the number of different models, saving costs and lowering prices.

The Broadband Market

Sometimes I wonder how much real demand there is for broadband services, and how much of the "demand" is really a push from the system operators and mobile device makers—the wireless version of super-sizing that fast-food meal.

I've decided that the present status is definitely a push from the providers. Sure, there is significant demand for high speed Internet with the convenience of cell phone style portability, but that demand is probably not sufficient to pay back the cost of initial deployment of such a major system upgrade. Thus, it is a business necessity to develop a consumer market in addition to a serious user market. Wireless operators are working hard to recruit casual users who are willing to pay for broadband access that is primarily used for personal entertainment.

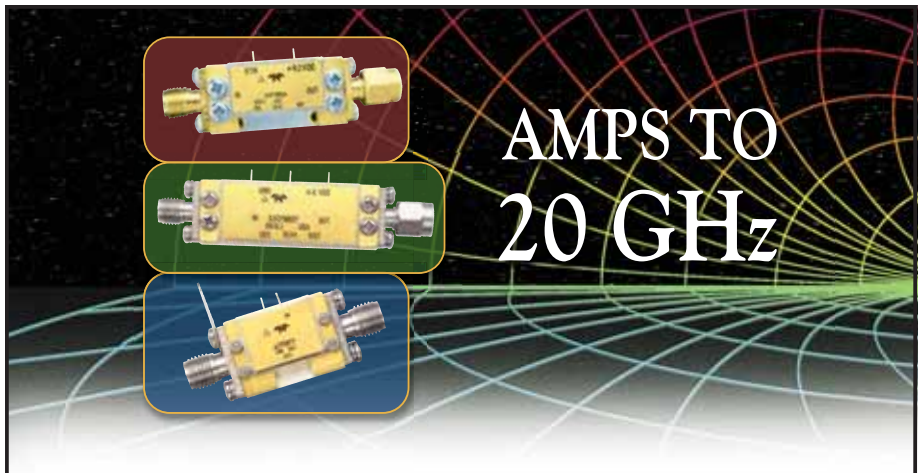
There is no doubt that the future will unfold with a vast array of services that take advantage of high speed wireless communications. And there is also no doubt that present standards used to start down that path will regularly be replaced with improved technologies.

Notes on Other Topics

Last issue's "Design Notes" column on writing styles and proper presentation of numbers and units generated plenty of comments. We received nearly as much feedback as the "Why 50 Ohms?" topic of a couple years ago! With so much feedback, plus additional information we have found, we decided to follow up on this topic in this month's Design Notes column.

The economic downturn seems to have created extra time for engineers to do some writing. Over the past few months, we have seen an increase in article proposals, particularly those with significant depth and substance. In this issue, Dr. Grebennikov's article on linearization and the review of 3G nonlinearity specs by engineers at Broadcom are good examples.

You can look forward to reading future articles on locating antennas within enclosures, receiver architecture planning, low voltage power amplifiers, software-defined radio, and an interesting examination of the combined analytical/intuitive process that engineers use to find new design solutions.



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ACP12019	6.0-12.0	9.5	4.0	28.0	39/52	10	210
A2CP14639	6.0-14.0	11.0	4.0	33.0	42/57	15	1500
ACP16025	8.0-16.0	7.5	4.3	28.0	42/65	12	253
ACP18015	8.0-18.0	9.0	4.0	15.5	23/31	5	63
A2CP18225	10.0-18.0	15.0	4.5	25.5	35/44	12	325
ACP20015	2.0-20.0	10.0	4.5	16.0	26/29	5	76
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Information: Conference Web site
http://www.wcai.com

September 21-25, 2009
IRMMW-THz 2009—The 34th International Conference on Infrared, Millimeter and Terahertz Waves
Seoul, Korea
Information: Conference Web site
http://www.irmmw-thz2009.org

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European Microwave Week
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Information: Conference Web site
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Information: Conference Web site
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October 6-8, 2009
RF & HYPER Europe 2009
Paris-Nord Villepinte, France
Information: Conference Web site
http://www.rfhyper.com

October 11-14, 2009
2009 IEEE Compound Semiconductor IC Symposium
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Information: Conference Web site
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October 27-29, 2009
3rd IEEE International Symposium on Microwave, Antenna, Propagation, and EMC Technologies for Wireless Communications
Beijing, China
Information: Conference Web site
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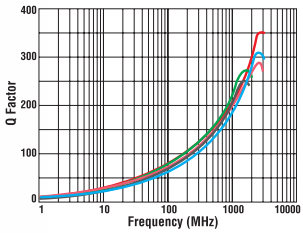

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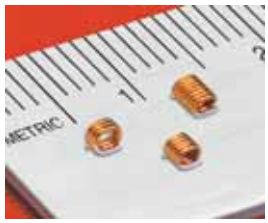

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Business News

ASK-intTag® LLC has signed an agreement to lease space at an IBM facility in Essex Junction, Vt. The company will occupy approximately 15,000 square feet of a separate and secured space at the IBM site to produce contactless inlays for a range of markets involving secure identification, including mass transit, bankcards, food, healthcare, gaming, and event ticketing. The company will be the first contactless inlay manufacturer in the United States to print silver-based antennas on inlays, the composition of which is highly resistant to forgery.

Aeroflex announced it is expanding its communications test capabilities with the strategic acquisition of **VI Technology, Inc.,** a provider of integrated baseband test solutions, specializing in audio, video and multi-media test.

Agilent Technologies Inc. introduced a new range of integrated curriculum-based solution kits for educators teaching analog-electronics, digital-systems and RF-communication classes. This further expands Agilent's range of education kits, which include solutions for basic electronics, RF circuit design and digital systems. Each solution includes carefully designed undergraduate engineering courseware, a mix of teaching slide sets, problem-based student labs and industry-standard measurement equipment.

Symmetricom®, Inc., a worldwide leader in precise time and frequency technologies that accelerate the deployment and enable the management of next generation networks, announced the company has been awarded a contract from **ITT Corporation Space Systems Division** valued at a minimum of $4 million, part of the building of GPS IIIA, the next generation Global Positioning System Space System program. Under the award, Symmetricom will deliver its Model 9552 Ovenized Quartz Master Oscillators over three years, including developing, qualifying and delivering flight oscillators for the first two space vehicles.

ITT Interconnect Solutions has partnered with the **Indian Space Research Organization** (ISRO) in the use of its Micro-D microminiature connectors for India's Chandrayaan-1 spacecraft, which is currently in lunar orbit. ITT's high-reliability Micro-D connectors are used to transmit information on data input and conversion equipment on the spacecraft, which is carrying, among other equipment, a state of the art imaging spectrometer that will provide the first map of the entire lunar surface at high spatial and spectral resolution.

Cirtronics Corporation is proud to celebrate their 30-year anniversary as a leader in turnkey electronic manufacturing services (EMS) with the expansion of their manufacturing facility in Milford, New Hampshire. An additional 75,000 square feet has been taken on since the beginning of the year to accommodate future growth and existing customers requiring high level PCB assembly, box-builds and testing. Cirtronics has maintained steady growth and now occupies almost the entirety of the Milford Technology Center.

TriQuint Semiconductor announced it has signed a framework agreement with **ZTE Corporation,** a manufacturer in China for wireless communication system equipment. The document states ZTE will procure TriQuint components, including but not limited to CDMA, GSM, and WCDMA ASICs, in the value of no less than US$50 million in the 2009 calendar year.

Azimuth Systems, Inc. announced that the company has partnered with the **Center for Multimedia Communication,** home of the Wireless Open-Access Research Platform (WARP) Group at Rice University. Azimuth will provide Rice University with an ACE™ channel emulator to enable extensive research on advanced transceiver architectures including beamforming, sphere detection and cooperative communications, to be conducted during the summer of 2009. In addition to allowing direct testing capabilities for MIMO projects, Azimuth and Rice University plan to collaborate on several research projects, which are poised to become part of 4G wireless.

Innovision Research and Technology plc and **NXP Semiconductors,** announced a joint marketing and licensing agreement related to the Near Field Communication technology. The deal brings together advanced Innovision IP and proven NXP expertise in firmware, software and secure elements, such as the SmartMX. With this combination both companies target to win NFC integration deals into deep sub micron based connectivity solutions. The agreement comes at a key point in the technology's development.

Marketplace News

Advances in semiconductor technology, favorable spectrum policy and demand for gigabit throughput capabilities have created an opportunity for millimeter wave radio technology to prosper. Applications in the mobile backhaul and consumer electronics markets in particular will drive the creation of an over $1.5 billion dollar market for upper millimeter-based technology in 2014, up from nearly $50 million in sales today, according to a new **Visant Strategies** report (www.visantstrategies.com). One of the best emerging markets for upper millimeter radio vendors is backhaul on the E-band for growing LTE, WiMAX, and HSPA+ wireless networks, according to report findings. Nearly 50 times the number of 60 GHz and E-Band point-to-point radios are due to be shipped in 2014 compared to 2008, according to "Upper Millimeter Microwave Radio: Ready to Compete 2009." The study examines the 60 GHz, E, Q and W bands and evaluates systems above 100 GHz. Existing and future applications such as mobile backhaul, wireless enterprise bridges, wireless fiber lateral emulations and government and public safety networks are studied. Also examined is use

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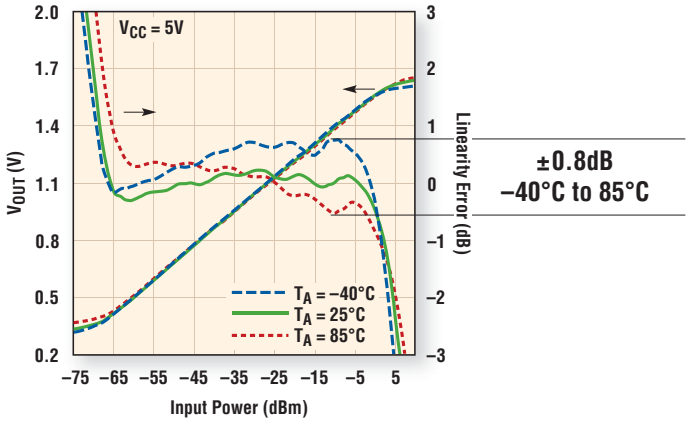

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	LT5581	40dB	10MHz to 6GHz	1.4mA @ 3.3V	3mm x 2mm DFN
Schottky Peak	LTC®5505	34dB	0.3GHz to 3GHz	0.5mA @ 3.3V	SOT-23
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within automotive collision avoidance, security/airport imaging applications and within personal area networks, particularly within WirelessHD.

The global base transmitting station (BTS) transceiver (TRx) market grew by 20.5% in units in 2008, according to the latest report from **EJL Wireless Research** titled "Global BTS Transceiver Market Analysis and Forecast, 5thEdition, 2008-2013."

"The transceiver market continued to show strong growth in 2008 despite the global financial crisis that emerged during the second half of the year and impacted wireless cap ex spending," says founder and President, Earl Lum. The report provides a unique perspective on the global shipments and demand for TRx equipment covering all air interface standards and frequencies and major OEMs including Alcatel-Lucent, Ericsson, Huawei Technologies, Motorola, Nokia Siemens Networks, Nortel Networks, and ZTE.

"The big surprise was Huawei Technologies which became the second largest vendor after Ericsson. In addition, ZTE moved into fourth place in overall rankings after posting 109% growth," says Lum. The overall market is expected to decline in 2009 and remain weak through 2011 before seeing a rebound in 2012. GSM shipments continue to represent the majority of the overall market demand in the forecast but will see a gradual decline. We forecast that LTE shipments will become the second largest segment by 2013, followed by W-CDMA/HSPA shipments.

The report is currently available for purchase and information can be downloaded at www.ejlwireless.com.

People in the News

TRAK Microwave Corporation, a **Smiths Interconnect** business, appoints **Keith Morrison** as Director, Business Development for Integrated Microwave Assemblies (IMA). He reports to Michael Kujawa, Vice President Sales and Marketing of TRAK Microwave Corporation. Mr. Morrison joins TRAK Microwave from Avnet, Inc. where his last position was Business Development Manager. Keith earned his BSEE from the University of South Florida, and brings to TRAK Microwave a successful 23-year career in the RF/Microwave industry, with proven skills in both technical and sales responsibilities. Keith and his family reside in the Tampa area, where he is active in boating, fishing, and coaching youth sports.

Sudhakar K. Rao, an internationally recognized expert in satellite antenna modeling, multiple-beam antennas and reconfigurable antenna payloads, is being honored by **IEEE** with the 2009 Judith A. Resnik Award. The award, sponsored by the IEEE Aerospace and Electronic Systems, IEEE Control Systems and IEEE Engineering in Medicine and Biology Societies, recognizes Rao for innovations and contributions leading to advanced multiple-beam and reconfigurable-beam payloads for communication satellites. Rao's pioneering work has resulted in advanced satellite payloads featuring increased capacity, improved performance and lower cost. Most of his inventions have already been deployed in space or are under preparation for deployment. An IEEE Fellow, Rao holds 32 U.S. patents and has published 130 technical papers. Rao is currently a corporate senior fellow at Lockheed Martin Space Systems, Newtown, Pa., where he has worked since 2003.

LadyBug Technologies LLC announced the appointment of **Paul Schmitz** as the company's Vice President of Sales and Marketing. In his new position Mr. Schmitz will hold responsibility for LadyBug's overall sales and marketing strategies worldwide, and will oversee the company's direct sales force, field representatives, and distribution network. He reports to LadyBug President Richard Hawkins. Mr. Schmitz brings three decades of high-technology product, marketing, and sales-management experience to his new position. Prior to joining LadyBug, he served as Senior Product/Platform Manager for Agilent Technologies' Microwave Signal Generator business sector, Santa Rosa, Calif. There he initiated successful new-product introductions, technical upgrades, and competitive-positioning enhancements resulting in both sales and market-share growth. Mr. Schmitz holds an M.S. in Engineering Management from Stanford University, Palo Alto, Calif., and a B.S. in Electrical Engineering from the University of Colorado, Boulder, Colo. He and his family reside in Forestville, Calif.

Alen Fejzuli, Conference Chair of **WAMICON 2009,** recently announced this year's **awards recipients** sponsored by CREE, Inc. at WAMICON 2009 in the following categories: Best Conference Paper: "A Multilayer Jerusalem Cross Frequency Selective Surface with Adequate Angular Stability at the 2.4 GHz ISM Band," S. Melais, T.Weller. Best Student Paper: "Study of a Polar Sigma-Delta Transmitter Associated to a High efficiency Switched Mode Power Amplifier for Mobile WiMax," F.Robert, M. Suarez PeÃloza, A. Diet, M. Villegas, G. Baudoin. Best Conference Poster: "Designing High Impedance Etched Ground GCPW," D. Schlieter. Best Student Poster: "Barium Strontium Titanate (BST) Microwave Devices," Tony Price, University of South Florida.

Sales Appointments

TT electronics IRC will represent **Semelab, Inc.** for North American sales and distribution of Semelab's complete product offering. Acquired by TT electronics in August 2008, Semelab designs, packages, tests and supplies discrete semiconductors, power modules, RF MOSFETs and sensor products to the high reliability, aerospace, medical and communication markets.

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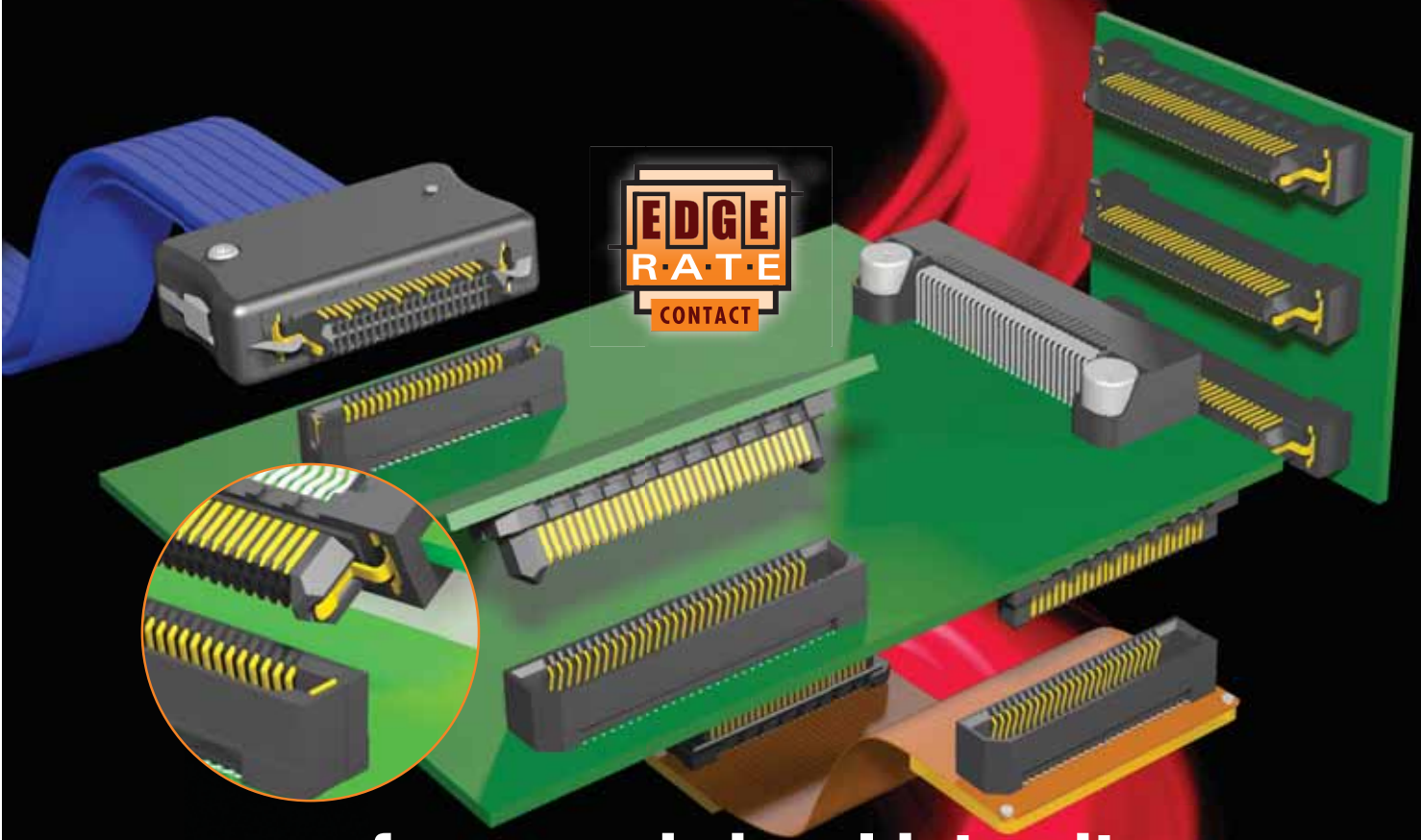

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IP2 and IP3 Nonlinearity Specifications for 3G/WCDMA Receivers

By Chris W. Liu and Morten Damgaard
Broadcom Corporation

Here is a thorough summary of linearity requirements for WCDMA receivers, with special attention to the way they apply to the direct conversion architecture

In 3G/WCDMA mobile handsets, the direct conversion receiver (DCR) architecture is commonly used due to its simplicity and low cost, where interstage filters are completely eliminated. However, with less filtering, the linearity requirement for a direct conversion receiver is very critical. It is essential to define system linearity requirements properly for the receiver to meet the performance requirements.

In FDD mode, the transmitter and receiver are continuously working at the same time, and the transmitted signal leaks into the receiver due to the limited TX-to-RX isolation of the duplexer. The transmitter is possibly the strongest interferer for the companion receiver in the handset, and poses the most stringent linearity requirement for the receiver.

I. 3GPP Test Cases

A number of test cases are specified in 3GPP for a WCDMA receiver, and each test case has different test conditions. Therefore nonlinearity performance requirements are needed of the receiver in each case. The related test cases are summarized in Tables A-F and used throughout the following discussion.

These test cases require the receiver to meet certain BER performance with these defined interferers. For RF system design, the requirements have to be interpreted into RF parameters such as noise figure (NF), compression point (P_{1dB}), second order intercept point (IP2), third order intercept point (IP3), and so forth.

	Test Conditions	
Band	DPCH_Ec	<REF$\hat{I}$or>
I	-117	-106.7
II	-115	-104.7
III	-114	-103.7
V	-115	-104.7
VIII	-114	-103.7
TX power: 24 dBm		

Table A · Sensitivity requirement.

	Test Conditions	
Parameter	**Case 1**	**Case 2**
DPCH_Ec	<REFSENS> + 14 dB	<REFSENS> + 41 dB
$\hat{I}$or	<REF$\hat{I}$or> + 14 dB	<REF$\hat{I}$or> + 41 dB
I_{oac}	-52	-25
F_{uw}	+5 or -5	+5 or -5
Tx power: 20 dBm		

Table B · Adjacent channel blocking.

II. Second Order Nonlinearity

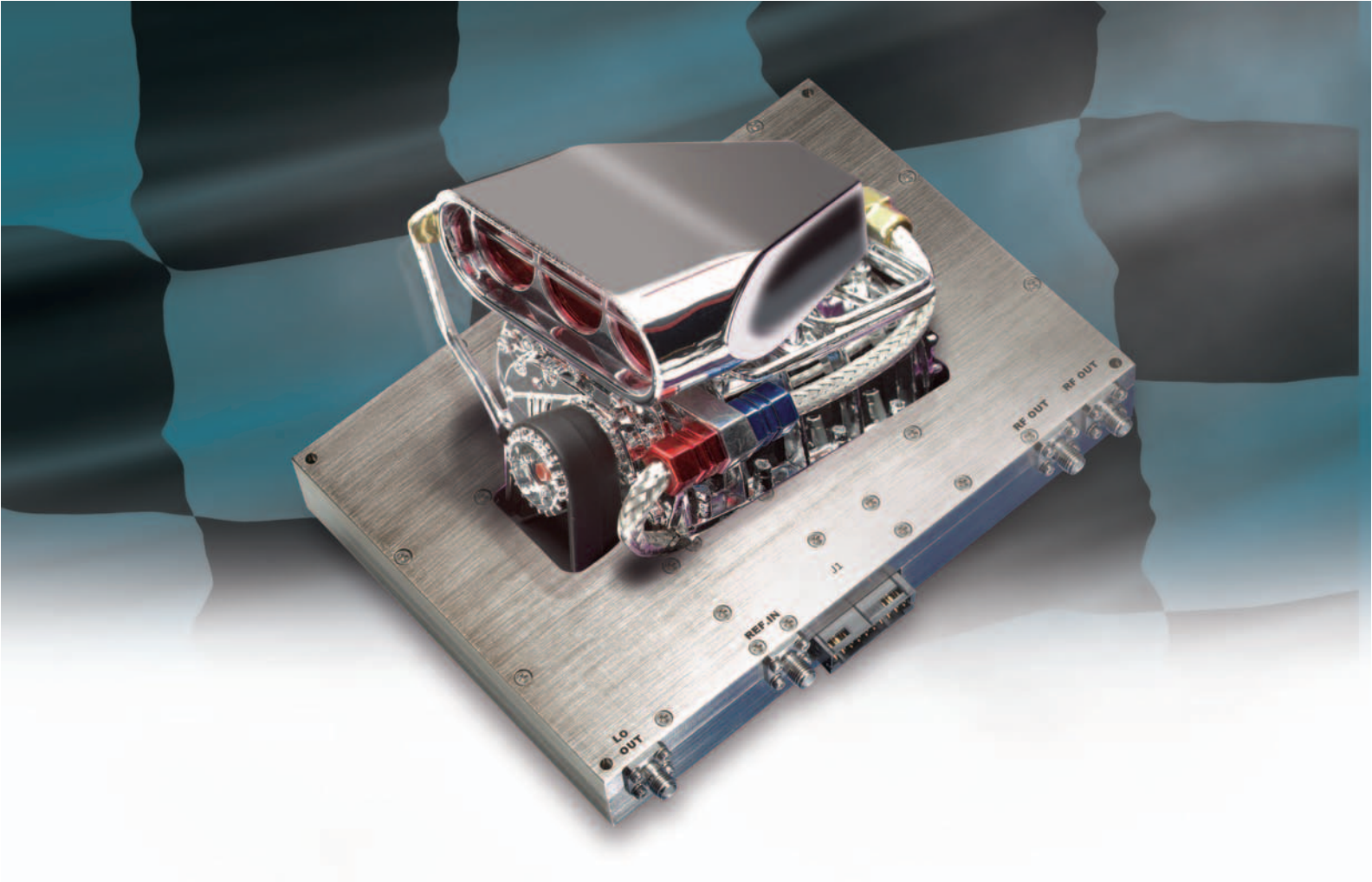
The second order nonlinearity of the receiver will square the modulated blocker signal, such as the TX leakage signal, producing DC and low frequency components which fall into the receive band of the direct conversion receiver. The AM (amplitude modulated) signal is demodulated into the RX channel with twice bandwidth of the original interferers. Moreover, a strong blocking signal will also intermodulate due to second order nonlinearity with the TX leakage signal to create a TX image which can fall into the band.

Mathematic Formula of IIP2

In general, the output signal of a nonlinear system can be described as follows:

$$V_o = a_1 V_i + a_2 V_i^2 + a_3 V_i^3 + \ldots$$

where V_o is the output voltage and V_i is the



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Wideband		Narrow band	
Parameter	Test Conditions	Band II, V	Band III, VIII
DPCH_Ec	<REFSENS> + 3 dB	<REFSENS> + 10 dB	
Îor	<REFÎor> + 3 dB	<REFÎor> + 10 dB	
I_{ouw1} (CW)	-46	-44	-43
I_{ouw2} mean power (modulated)	-46	-44	-43
F_{uw1} (offset)	± 10	± 3.5	± 3.6
F_{uw2} (offset)	± 20	± 5.9	± 6.0
Tx power: 20 dBm			

Table C · Intermodulation (wideband and narrowband).

Parameter	Unit	Test Conditions	
		Band II, V	Band III, VIII
DPCH_Ec	dBm/3.84	<REFSENS> + 10 dB	
Îor	dBm/3.84	<REFÎor> + 10 dB	
$I_{blocking}$	dBm	-57	-56
F_{uw} offset	MHz	± 2.7	± 2.8
Tx power: 20 dBm			

Table E · Narrow band blocking (NBB).

Parameter	Unit	Freq range 1	Freq range 2	Freq range 3
DPCH_Ec	dBm/3.84	<REFSENS>+3 dB	<REFSENS>+3 dB	<REFSENS>+3 dB
$\hat{I}_{or}$	dBm/3.84	<REFÎor> + 3 dB	<REFÎor> + 3 dB	<REFÎor> + 3 dB
$I_{blocking}$ (CW)	dBm	-44	-30	-15
F_{uw} (Band I)	MHz	2050<f <2095 2185<f <2230	2025 <f ≤2050 2230 ≤f <2255	1< f ≤2025 2255≤f<12750
Tx power: 20 dBm				

Table D · Out of band blocker (OOB).

Parameter	Unit	Test Conditions	
DPCH_Ec	dBm/3.84	<REFSENS> + 3 dB	
Îor	dBm/3.84	<REFÎor> + 3 dB	
$I_{blocking}$	dBm	-56	-44
F_{uw} offset	MHz	± 10	± 15
Tx power: 20 dBm			

Table F · In band blocking (IBB).

input voltage.

If using a traditional two tone signal as input V_i, we have:

$$V_i = A \cdot \cos(\omega_1 t) + A \cdot \cos(\omega_2 t)$$

The second order component of the output is:

$$\begin{aligned} V_o &= a_2 V_i^2 \\ &= a_2 A^2 \left(\cos^2(\omega_1 t) + \cos^2(\omega_2 t) + 2\cos(\omega_1 t)\cos(\omega_2 t)\right) \\ &= a_2 A^2 \left(1 + \frac{1}{2}\cos(2\omega_1 t) + \frac{1}{2}\cos(2\omega_2 t) + \cos((\omega_1 + \omega_2)t) + \cos((\omega_1 - \omega_2)t)\right) \end{aligned}$$

It shows that the second order IM2 products are created at three frequencies: DC, $f1 + f2$ and $f1 - f2$. In terms of power level, IM2 products are distributed against total IM2 power as:

- 50% (–3 dB) at DC
- 25% (–6 dB) at $f1 + f2$
- 25% (–6 dB) at $f1 - f2$

The IM2 of low frequency is considered only in this paper since it is the one falling into band. The power level of the IM2 product at $f1 - f2$ is 25% of the total IM2 power which is 6 dB below the total IM2 power. So the power level of IM2 at low frequency ($f2 - f1$) can be expressed as:

$$P_{im2}(\text{dBm}) = 2 \cdot P_{in} - \text{IIP2} - 6 \text{ dB} \quad (1)$$

IIP2 With WCDMA TX Uplink Leakage

If the signal is AM modulated, such as TX leakage signal, the 2 tone formula above can not be fully applied. The difference between the formula and 2 tone signal is called correction factor [4]. The following section derives the correction factor for a WCDMA signal using ADS simulation (Fig. 1).

The two tone test case and the case of single tone plus WCDMA uplink signal test case are simulated. The coefficients of the second order product in the model is set to 0.1 for simplicity since only the difference between using the two tone measurement and the TX uplink modulated signal is needed. The two tone simulation results are shown in Figure 2 and the simulation results with TX uplink signal are shown in Figure 3.

With the TX uplink signal, since it is modulated, the low frequency IM2 products are measured by integrating the power from 1 kHz to 2.0 MHz in the frequency domain. The difference of IM2 between the two tone and modulated signal is 9.72 dB. The IP2 two-tone formula for TX leakage case with correction factor is as follows:

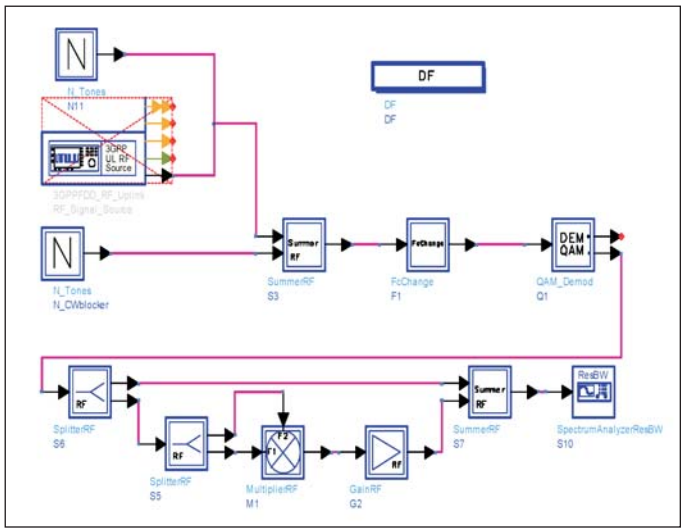

Figure 1 · Advanced Design System (ADS) bench for IP2 simulation.

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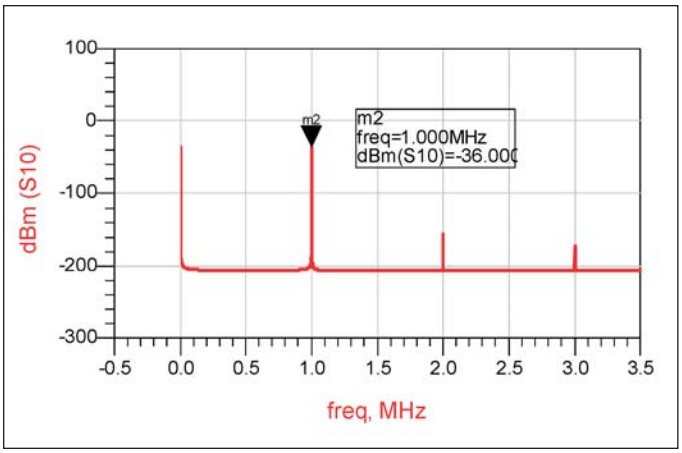

Figure 2 · Low frequency output of two tone simulation for IP2.

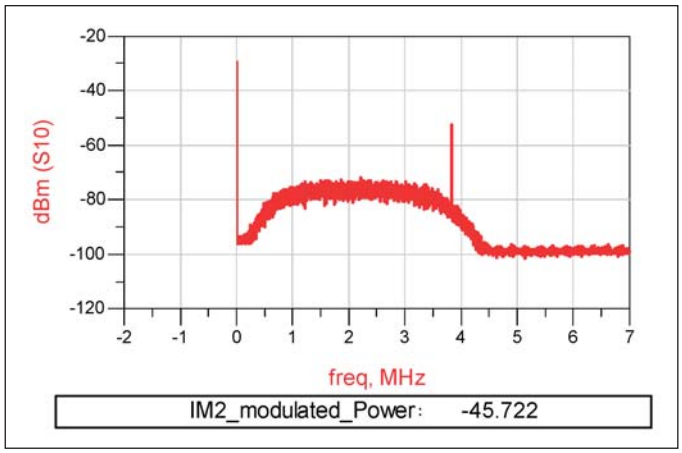

Figure 3 · Low frequency output of WCDMA uplink simulation for IP2.

$$P_{\text{im2\_txleakage}} = 2 \cdot P_{\text{in}} - \text{IIP2} - 15.722 \text{ dB} \quad (2)$$

IIP2 With In-Band Blocking

One of 3GPP test cases is adjacent blocker test, in which the modulated downlink blocking signal is injected into the receiver with offset frequency at ±5 MHz. By using method similar to the one discussed above, with an adjacent blocker signal (Test Model 1), the corrected formula for IIP2 with a WCDMA downlink signal is:

$$P_{\text{im2\_adj}} = 2 \cdot P_{2T} - \text{IIP2} - 3.87 \text{ dB} \quad (3)$$

IIP2 With Out of Band Blocker (OOB)

In OOB test cases, depending on the blocker frequency, the IM2 products of the OOB blocking signal and TX leakage signal may fall into the RX band. The power of IM2 can be calculated using the formula in Equation 4.

$$P_{\text{im2\_OOB}} = P_{\text{CW}} + P_{tx} - \text{IIP2} \quad (4)$$

III. Third Order Nonlinearity

For third order nonlinearity, the 3GPP intermodulation test case defines the IP3 requirements of the receiver. However the blockers such as adjacent channel blockers, narrow band blockers, out of band blockers, etc either leak into the RX channel, or cross-modulate, or intermodulate with the TX leakage signal, the distortion products falling into wanted channel. The receiver needs to have good linearity performance under all blocker conditions.

IIP3 With Two Tone Test

Assuming the input signal as V_i with two tone signals,

$$V_i = A1 \cdot \cos(\omega_1 t) + A2 \cdot \cos(\omega_2 t)$$

The output signal $y(t)$ can be expressed as:

$$y(t) = a_1 \cdot V_i(t) + a_2 \cdot V_i^2(t) + a_3 \cdot V_i^3(t)$$

The third order intermodulation products are:

$$y(t)_{\text{3rdorder}} = 3a_3A_1^2A_2\cos(\omega_1)^2 t\cos(\omega_2)t + 3a_3A_1A_2^2\cos(\omega_1)t\cos(\omega_2)^2 t = \frac{3a_3A_1^2A_2}{4}\cos(2\omega_1 - \omega_2)t + \frac{3a_3A_2^2A_1}{4}\cos(2\omega_2 - \omega_1)t + \ldots$$

The third order IM3 products are created at: $2f1 + f2$, $2f1 - f2$, $2f2 - f1$ and $2f2 + f1$. In a down conversion receiver, only the low frequency products are interesting. The power of IM3 at low frequency is expressed in following classical formula:

1. IM3 at $2f2 - f1$:

$$P_{\text{IIM3}}(\text{dBm}) = 2 \cdot P_2 + P_1 - 2 * \text{IIP3} \quad (5)$$

2. IM3 at $2f1 - f2$:

$$P_{\text{IIM3}}(\text{dBm}) = 2 \cdot P_1 + P_2 - 2 * \text{IIP3} \quad (6)$$

If two tones are equal, the formula becomes:

$$P_{\text{IIM3}}(\text{dBm}) = 3 \cdot P_{\text{in}} - 2 * \text{IIP3} \quad (7)$$

IIP3 With Cross Modulation

The TX signal leakage signal can be cross modulated by a strong blocker such as a narrow band blocker, adjacent channel blocker, with the cross modulated signal falling into the receive channel.

An example is illustrated in Figures 4 and 5. Figure 4 shows the output of a nonlinear circuit with two tone signal. Replacing one tone by a TX uplink leakage signal in

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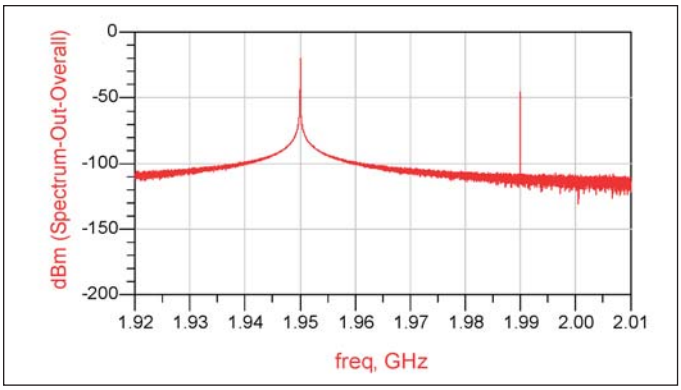

Figure 4 · Two tone simulation.

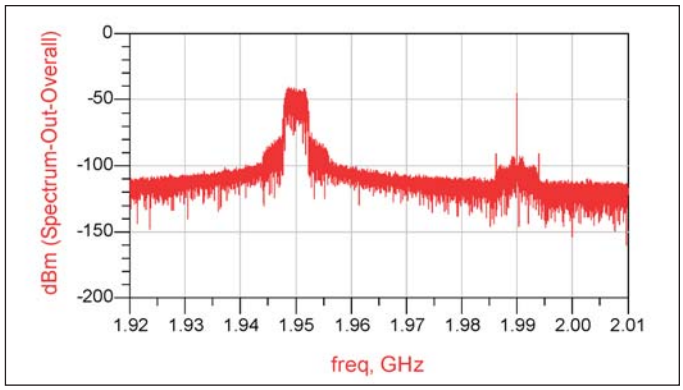

Figure 5 · One tone is replaced by TX leakage signal.

Figure 5, cross modulation products are observed around the single tone at 1.99 GHz.

For mathematical analysis, we could assume that the signal at the LNA input comprises of two interferers besides the wanted signal, namely a CW blocker signal and the TX leakage signal. Note that the TX leakage signal is amplitude modulated, while the blocker is not modulated in this analysis. Assuming the input signal $x(t)$, with blocker and TX leakage signal then we get:

$$x(t) = A_1 \cdot \cos(\omega_1 t) + A_2 \cdot [1 + m(t)] \cdot \cos(\omega_{tx} t)$$

where $m(t)$ is the amplitude modulation having a fundamental frequency at WCDMA chip rate.

When the signal inject into the non-linear RF circuit, receiver in this case, the output signal can be expressed as follows, considering only up to third order nonlinearity:

$$\begin{aligned} y(t) &= a_1 \cdot x(t) + a_2 \cdot x^2(t) + a_3 \cdot x^3(t) \\ &= a_1 A_1 \cdot \cos(\omega_1 t) + a_1 \cdot A_2 \cdot (1 + m(t)) \cos(\omega_1 t) \\ &+ a_2 \cdot A_1^2 \cdot \cos^2(\omega_1 t) + a_2 \cdot A_2^2 \cdot (1 + m(t))^2 \cos^2(\omega_{tx} t) \\ &+ 2a_2 \cdot (1 + m(t)) \cdot A_1 \cdot A_2^2 \cdot \cos(\omega_1 t) \cdot \cos(\omega_{tx} t) + a_3 A_1^3 \cos^3(\omega_1 t) \\ &+ 2a_3 \cdot A_1^2 \cdot A_2 \cdot \cos^2(\omega_1 t) \cdot \cos(\omega_{tx} t) \\ &+ a_3 \cdot A_1^2 \cdot A_2 \cdot (1 + m(t)) \cos^2(\omega_1 t) \cdot \cos(\omega_{tx} t) \\ &+ 3 \cdot a_3 \cdot A_1 \cdot A_2^2 \cdot (1 + m(t))^2 \cdot \cos(\omega_1 t) \cdot \cos^2(\omega_{tx} t) \end{aligned}$$

The term $3 \cdot a_3 \cdot A_1 \cdot A_2^2 \cdot (1 + m(t))^2 \cdot \cos(\omega_1 t) \cdot \cos^2(\omega_{tx} t)$ can be further expanded to:

$$\begin{aligned} &3/2 \cdot a_3 \cdot A_1 \cdot A_2^2 \cdot (1 + m(t))^2 \cdot \cos(\omega_1 t) \\ &+ 3/2 \cdot a_3 \cdot A_1 \cdot A_2^2 \cdot (1 + m(t))^2 \cdot \cos(\omega_1 t) \cos(2\omega_{tx} t) \end{aligned}$$

The term $3/2 \cdot a_3 \cdot A_1 \cdot A_2^2 \cdot (1 + m(t))^2 \cdot \cos(\omega_1 t)$ shows the blocker signal being modulated by the square of the amplitude of the TX leakage.

Cross Modulation With Adjacent Channel Blocker

The straight forward thinking is to use the IIP3 formula with little modification. The quantity of the cross-modulation product can be evaluated with the approximated formula:

$$P_{\text{crossmod}} = C_{\text{factor}} + 2P_{\text{TX}} + P_{\text{adj}} - 2\text{IIP3} \quad (8)$$

where C_{factor} is the correction factor that takes into account the difference between using two CW tone measurement and modulated signal measurement. The correction factor has been determined by comparing the difference between two tone case and the TX leakage case in simulation. It is found that C_{factor} is around 7.4 dB. The corrected formula for the adjacent channel blocker test case is:

$$P_{\text{crossmod\_adj}} = 2P_{\text{TX}} + P_{\text{adj}} - 2\text{IIP3} - 7.4 \quad (9)$$

Cross Modulation With Narrow Band Blocker

The narrow band blocker appears at ±2.7 (2.8) MHz, the offsets are much closer to the carrier compared the adjacent channel blocker case. The approximate formula is found to be:

$$P_{\text{crossmod\_nb}} = 2P_{\text{TX}} + P_{\text{nb}} - 2\text{IIP3} - 2.4 \quad (10)$$

Adjacent Channel Leakage—ACLR

The receiver front end nonlinearity can create spectrum re-growth for an adjacent channel blocker, partially falling into the wanted band. The approximate formula to calculate the leakage power is reported in [3] as:

$$P_{\text{aclr}} = -20.75 + 2 \cdot (P_{\text{in}} - \text{IIP3}) + 1.6 \cdot \text{PAR} \quad (11)$$

where PAR is the Peak to Average ratio of the downlink signal.

IV. System Specification

In order to determine the IP2 and IP3 requirements, it is essential to consider all impairments including IM2,

IM3 products, which must be low enough for the receiver sensitivity degradation to be acceptable. In the following sections, IP2 and IP3 are derived. The following analysis are applied to Band-I typical case.

IP2 at TX Frequency Requirement at Sensitivity Level

Maximum Allowed Noise

First of all, let's look at the maximum allowed noise floor without any interferers. From the 3GPP specification, the specified sensitivity is:

- Total input power: I_{or} = –106.7 dBm/3.84 MHz,
- Dedicated physical channel power: DPCH_EC = P_{sens} = –117 dBm/3.84 MHz.
- The processing gain of 12.2 kps reference channel: G_p = 25 dB

It has been reported that the required E_b/N_t, the ratio of signal energy bit to noise spectra density, shall be better than 5.2 dB to guarantee the Base Band (BB) modem can demodulate the WCDMA signal properly. In this paper, we assume E_b/N_t as 5.2 dB and add 2.2 dB implementation margin. Therefore, the allowed total noise power at antenna input is:

$$P_n = P_{sens} + G_p - \frac{E_b}{N_t} = -117 + 25 + 7.4 = -99.42 \text{ dBm}$$

Typically some margin is needed to cover process variation for production yield. By applying 1 dB margin, the total noise allowed at antenna connector is:

$$N_{\text{tmax\_ant}} = P_n - 1 = -100.4 \text{ dBm/3.84 MHz}$$

At RFIC input, taking front end loss into account, the maximum allowed interfere level is:

$$N_{\text{tmax}} = N_{\text{tmax\_ant}} - L_{\text{front}} = -104.2 \text{ dBm}$$

where L_{front} = 3.8 dB is assumed including switch loss and duplexer RX insertion loss, etc.

The distortion products due to component nonlinearity shall be kept below a certain level so that the receiver sensitivity will not be degraded too much. The interferers are not only from second and third order nonlinearity but also from other sources, which include:

- PA noise at RX band
- RFIC phase noise at RX band
- TX leakage IIP2
- TX leakage reciprocal mixing

Additionally, if blockers are present, other concerns also include:

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- Blocker reciprocal mixing
- Cross modulation
- Adjacent blocker—ACLR
- Front end switch IM2/IM3 products

TX Noise at RX Band

First of all, the typical performance of the PA, RFIC and duplexer is studied in this case. Though the components from different manufacturers could perform differently, the typical performances used in this paper are summarized in Tables 1, 2 and 3, which are based on major manufacturers' datasheets.

It should be noted that for the sensitivity test case, the transmitter output power is defined as maximum output power, 24 dBm, at the antenna connector. In all other test cases, the output power at the antenna connector is defined as 20 dBm according 3GPP specifications.

The output power of the PA can be calculated by add loss between the PA output and the antenna connector.

$$PA_{\text{out}} = P_{\max} + FE_{\text{loss}} + Dup_{\text{loss}}$$

where FE_{loss} is the front end loss, including switch, PCB traces and matching loss. Dup_{loss} is the duplexer TX insertion loss.

The RFIC output power can be calculated based on PA_{out} and the PA gain. It is around 0.5 dBm, P_{rfic} = 0.5 dBm, in this case.

The noise present in the RX band at the LNA input due to PA phase noise is

$$P_{\text{pa\_noise}} = PA_{\text{noise}} - \text{ISO}_{\text{tx–rx}} + 10^{*}\text{Log}_{10}(\text{BW}) = -118.15\ \text{dBm}$$

where $\text{ISO}_{\text{tx–rx}}$ is duplexer isolation from TX to RX at RX band, and channel bandwidth BW = 3.84 MHz.

And the TX noise at RX band due to RFIC is:

$$P_{\text{rfic-noise}} = P_{\text{n\_rfic}} + PA_{\text{gain}} + 10^{*}\text{Log}_{10}(\text{BW}) - \text{ISO}_{\text{tx–rx}}$$

$$= -120.15\ \text{dBm}$$

Since the total noise due to TX transmitter is the summation of both, it becomes:

$$P_{\text{tx\_noise}} = 10\log 10\left(10^{\frac{PA\text{tx\_noise}}{10}} + 10^{\frac{\text{Pr fic\_noise}}{10}}\right) = -114.86\ \text{dBm}$$

TX Leakage Reciprocal Mixing

The TX leakage power at the RFIC input is:

$$P_{\text{leakage}} = PA_{\text{out}} - \text{ISO}_{\text{tx\_rx}} = -25.6\ \text{dBm}$$

Assuming the phase noise of the RX LO at the TX frequency is –157 dBc/Hz, then the reciprocal mixing at the RX band is:

$$P_{\text{tx\_reciprocal}} = -116.75\ \text{dBm}$$

	Band	Gain	Rx Noise (dBm/Hz)
PA	1	27	-140
PA	2	27	-140
PA	5	27	-140

Table 1 · PA noise and gain.

	RX-ANT(dB)	TX-ANT(dB)		RX-TX (dB)	
Band	Insertion loss	Insertion loss	RX noise attenuation	TX attenuation	RX attenuation
1	2.3	1.9	45	53	44
2	3.9	3.2	45	53	48
5	2.7	2.3	45	50	45

Table 2 · Duplexer performance.

	Band	Rx Noise (dBc/Hz)
RFIC	1	-165
RFIC	2	-165
RFIC	5	-170

Table 3 · RFIC phase noise at RX band.

TX Leakage IIP2

First, let's make an assumption that the allowed sensitivity degradation as X dB, and then we will have:

$$10^{\left(\frac{PA\text{tx\_noise}}{10}+\frac{PA\text{tx\_im2}}{10}+\frac{Nt\ \max}{10}+\frac{P\text{tx\_reciprocal}}{10}\right)} = 10^{\frac{Nt\max+X}{10}}$$

Arranging the equation above and expressing the above equation in dB, we get:

$$10\log 10\left(10^{\frac{PA\text{tx\_noise}}{10}} + 10^{\frac{P\text{tx\_im2}}{10}} + 10^{\frac{P\text{tx\_reciprocal}}{10}}\right)$$

$$= N_{t\max} + 10\log 10\left(10^{\frac{X}{10}} - 1\right)$$

The term

$$10\log 10\left(10^{\frac{X}{10}} - 1\right)$$

is the factor that determines how much the impairments shall be below $N_{t\max}$. For example, if X = 0.3 dB, meaning 0.3 dB sensitivity degradation is allowed, the total impairment power level must be 11.4 dB below the maxi-

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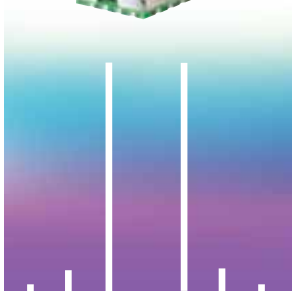

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TYPICAL SPECIFICATIONS

Model No.	Frequency (MHz) RF	LO	IF	LO Pwr. (dBm)	IP3 (dBm)	1dB Comp. (dBm)	Conv.Loss (dB)	Isolation (dB) L-R	L-I	Price $ ea. Qty.(1-9)
LAVI-9VH+	820-870	990-1040	120-220	+19	+36	+23	7.2	46	46	15.95
LAVI-10VH+	300-1000	525-1175	60-875	+21	+33	+20	6.3	50	45	22.95
LAVI-17VH+	470-1730	600-1800	70-1000	+21	+32	+20	6.8	52	50	22.95
LAVI-22VH+	425-2200	525-2400	100-700	+21	+31	+20	7.7	50	45	24.95
LAVI-2VH+	2-1100	2-1100	2-1000	+23	+34	+23	7.5	48	47	24.95
LAVI-25VH+	400-2500	650-2800	70-1500	+23	+32	+20	7.5	50	45	24.95

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mum allowed noise floor $N_{t\max}$. In general, it is good to keep IM2 due to TX leakage between 13 to 16 dB below $N_{t\max}$ so that the IM2 product degrades the sensitivity by only 0.1 to 0.3 dB. In this paper 16 dB is chosen for better performance. If so, P_{im2} shall be less than –120 dBm that is 16 dB below $N_{t\max}$.

From equation (2), IIP2 at the TX frequency can be derived as:

$$\text{IIP2(dBm)} = 2 \cdot P_{2T}\text{(dBm)} - P_{\text{im2\_tx}} - 15.722\ \text{dB} = 48\ \text{dBm}$$

Therefore the IIP2 requirement due to TX leakage for Band I at the RFIC input should be better than 48 dBm at the TX frequency. With IIP2 = 48 dBm, adding all other impairments, the overall sensitivity degradation is less than 0.6 dB. Further improvements in the PA phase noise and duplexer attenuation will continue to minimize the degradation.

Adjacent Channel Blocker Case 1

The adjacent blocker appears at 5 MHz offset with a power level of –52 dBm at antenna connector. Since the input wanted signal is 14 dB higher than sensitivity, it is reasonable to assume that the degradation due to the total interferer power shall be less than 14 dB. However in practice, margin is needed.

Note that with an adjacent channel blocker, the major contributor of in channel distortion is from ACLR, rather than IP2. First of all, we need to determine the power leakage into the channel due to adjacent channel blocker, P_{aclr}. To minimize the impact of ACLR, in this paper, P_{aclr} is kept 16 dB below the wanted signal.

The allowed total noise floor N_t, is 14 dB higher than $N_{t\max}$ in this test case, therefore N_t = –90.2 dBm, The allowed P_{aclr} will be $P_{\text{aclr}} = N_t$ –16 = –106.2 dBm. From the equation (10), IIP3 can be calculated,

$$\text{IIP3} = \frac{1}{2}\left(-P_{\text{aclr}} - 20.75 + 2 \cdot P_{\text{adj}} + 1.6 \cdot PAR\right) = -7.2\ \text{dBm}$$

We also need to check the IP3 requirements due to Cross modulation. Based on the cross modulation formula derived in previous section, equation (8), the cross-modulation product can be calculated:

$$P_{\text{crossmod}} = 2P_{\text{TX}} + P_{\text{adj}} - 2\text{IIP3} - 7.4 = -108.27\ \text{dBm}$$

The cross modulation product with IP3 = –7.2 dB is 18 dB below the N_t, and not a problem.

Then $P_{\text{im2\_adj}}$ due to adjacent channel blocker can be assumed 16 dB below the N_t as well. Then IIP2 at 5 MHz adjacent channel, IIP2_adj is obtained:

$$\text{IIP2(dBm)} = -P_{\text{IIM2\_adj}} + 2 \cdot P_{\text{adj}} - 3.87\ \text{dB} = -15\ \text{dBm}$$

Adjacent Channel Case 2

The blocker is present at 5 MHz offset with a power level of –25 dBm and the input wanted signal is 41 dB higher than sensitivity. In this case, N_t = –63.2 dBm, the allowed P_{aclr} is:

$$P_{\text{aclr}} = N_t - 16 = -79.2\ \text{dBm}.$$

The IIP3 and cross modulation product are calculated below.

$$\text{IIP3} = \frac{1}{2}\left(-P_{\text{aclr}} - 20.75 + 2 \cdot P_{\text{adj}} + 1.6 \cdot \text{PAR}\right) = 6.2\ \text{dBm}$$

$$P_{\text{crossmod}} = 2P_{\text{TX}} + P_{\text{adj}} - 2\text{IIP3} - 9.67 = -110.7\ \text{dBm}$$

Similar to the adjacent channel blocker case 1, IIP2 can be derived for the adjacent channel blocking case 2:

$$\text{IIP2(dBm)} = -P_{\text{IIM2\_adj}} + 2 \cdot P_{\text{adj}} - 3.87\ \text{dB} = 12\ \text{dBm}$$

In-Band Blocker 10 MHz and 15 MHz

The input wanted signal is 3 dB higher than sensitivity and the blocker at 10 MHz offset has a power level at –56 dBm, and –44 dBm at 15 MHz offset. Since the offset frequency is far from the carrier, there is no concern for ACLR in this case.

Using equation (3), keeping the IM2 product 16 dB below N_t, IIP2_adj at 10 MHz and 15 MHz are computed:

$$\text{IIP2}_{10\ \text{MHz}} = -P_{\text{IIM2\_10 MHz}} + 2 \cdot P_{10\ \text{MHz}} - 3.87\ \text{dB} = -12\ \text{dBm}$$

$$\text{IIP2}_{15\ \text{MHz}} = -P_{\text{IIM2\_15 MHz}} + 2 \cdot P_{15\ \text{MHz}} - 3.87\ \text{dB} = 12\ \text{dBm}$$

Intermodulation

For band I, the narrow band blocking test case does not apply, but for other bands. It does in the wideband intermodulation test case, one of the blocking signals is a CW tone at –46 dBm, while another blocker is a WCDMA modulated signal with power level of –46 dBm. The wanted signal is 3 dB higher than sensitivity power level.

Keeping the IM3 product 16 dB below the N_t, which is 3 dB above $N_{t\max}$, we can compute the IIP3 using equation (7).

$$\text{IIP3} = 0.5 * (3 \cdot P_{\text{in}} - P_{\text{IIM3}}) = -15\ \text{dBm}$$

Out of Band Blocker

For out of band (OOB) blocking cases, we need to consider both the second order and third order distortion from the switches that is between the duplexer and the antenna. The switches in the front end module generate second and third order nonlinearity components falling in RX band as well. Bases on the manufacturer's datasheet, it is assumed in this paper that IM2 and IM3 of the front

end switch module are:

P_{sw_im2} = –110 dBm
P_{sw_im3} = –110 dBm

With the OOB test case, not all blocker frequencies produce distortion that falls into wanted band. For the second order nonlinearity, we will consider:

- $F_{cw} = F_{rx} - F_{tx}$
- $F_{cw} = F_{rx} + F_{tx}$

For the third order nonlinearity, we will consider:

- $F_{cw} = F_{rx} - \Delta F/2$
- $F_{cw} = F_{rx} - 2*\Delta F$
- $F_{cw} = 2* F_{tx} + Frx$
- $F_{cw} = \Delta F/2$

where F_{cw} is the blocker frequency, F_{rx} is the RX channel frequency, F_{tx} is the TX frequency related to F_{rx}, and ΔF is the frequency separation between F_{rx} and F_{tx}.

$F_{rx} - F_{tx}$

The CW blocking signal is at very low frequency, which is the separation between RX and TX frequencies. The second order IM2 results in mixing of F_{tx}+ ($F_{rx} - F_{tx}$) = F_{rx}.

As the CW blocker level is –15 dBm, and the duplexer attenuation at the CW blocker frequency is around 30 dB for Band I, we get P_{cw} = –45 dBm at the input of the RFIC and the TX leakage power level P_{tx} = –29.6 dBm.

Using equation (4), with P_{im2} 16 dB below N_t, which is 3 dB above noise floor of –104.2 dBm, IIP2 is calculated as:

$$IIP2 = P_{CW} + P_{TX} - P_{IM2} = 38 \text{ dBm}$$

$F_{tx} + F_{rx}$

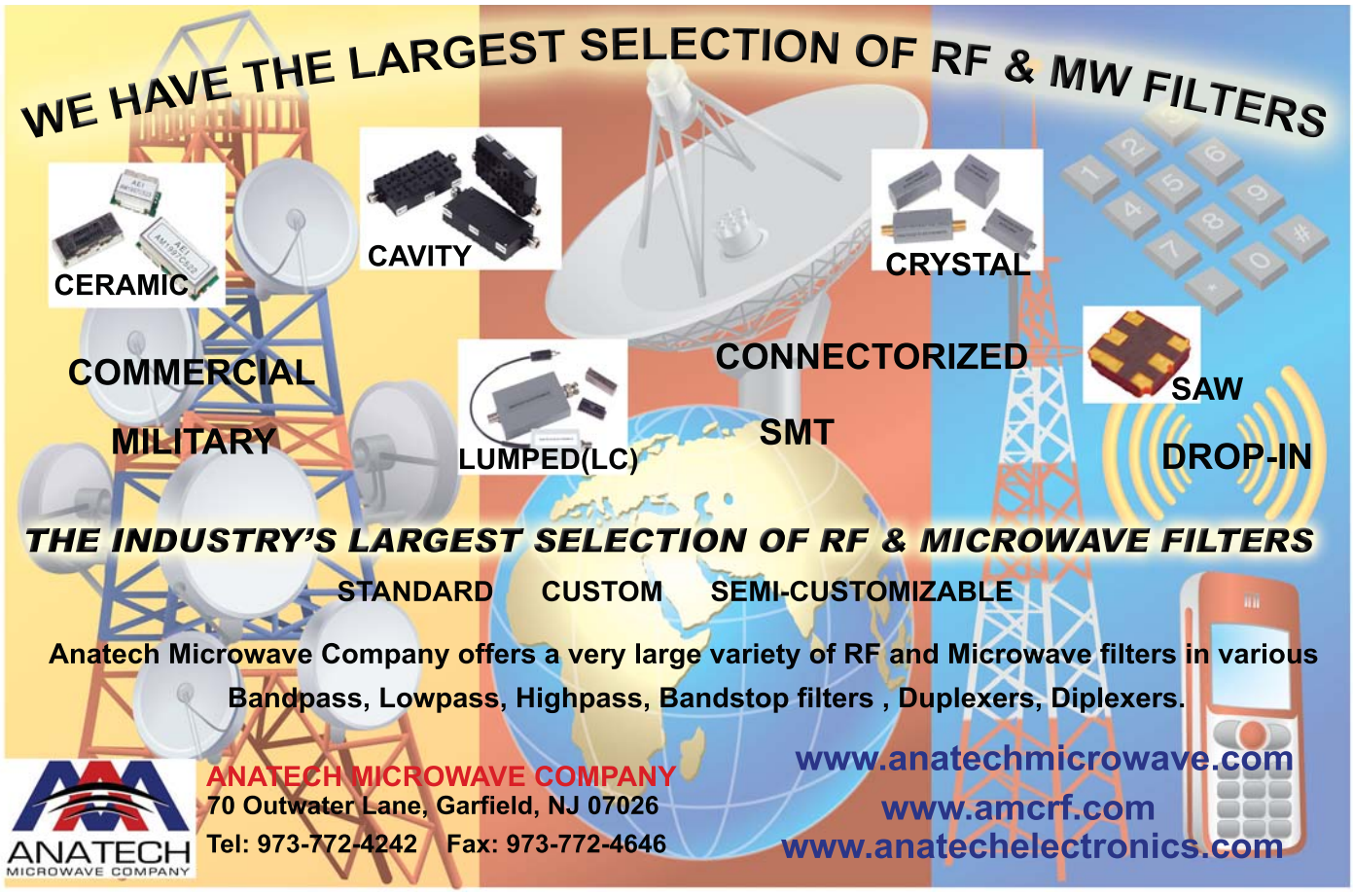
In this case, the OOB signal is at high frequency. The mixing of ($F_{tx} + F_{rx}$) – F_{tx} = F_{rx}, falling in the RX band. The blocker level is –15 dBm, and the duplexer attenuation at the blocker frequency is around 30 dB for Band I so that P_{cw} = –45 dBm, and P_{tx} = –29.6 dBm. Using equation (3), with P_{im2} 16 dB below N_t, we have

$$IIP2 = P_{CW} + P_{TX} - P_{IM2} = 38 \text{ dBm}$$

$F_{rx} - \Delta F/2$

The blocker power level actually varies depending on frequency offset according the 3GPP requirements.

When F_{cw} is between 2050 MHz and 2075 MHz, the blocker level is –44 dBm. As the duplexer attenuation at



blocker frequency is around 5 dB, P_{cw} is –49 dBm. IIP3 is calculated as:

$$IIP3 = 0.5(2 \cdot P_{CW} + P_{TX} - P_{im3}) = -7 \text{ dBm}$$

When F_{cw} is located between 2025 to 2050 MHz, the blocker power level is –30 dBm. Assuming duplexer attenuation at blocker frequency is 15 dB, P_{cw} is –45 dBm. IIP3 is calculated as:

$$IIP3 = 0.5(2 \cdot P_{CW} + P_{TX} - P_{im3}) = -3 \text{ dBm}$$

When F_{cw} is between 2015 and 2015 MHz, the blocker power level is –15 dBm. With the duplexer attenuation at the blocker frequency of 30 dB, P_{cw} is –45 dBm. IIP3 is calculated as:

$$IIP3 = 0.5(2 \cdot P_{CW} + P_{TX} - P_{im3}) = -3 \text{ dBm}$$

F_{rx} – 2*ΔF

The CW blocker frequency is located between 1730 and 1780 MHz. To allow P_{im3} = –117 dBm, which is 16 dB below the maximum allowed noise floor N_t, the blocker power level is –15 dBm. Assuming duplexer attenuation at the blocker frequency is 38 dB, P_{cw} is –54 dBm. IIP3 is calculated as:

$$IIP3 = 0.5(P_{CW} + 2P_{TX} - P_{im3}) = 1 \text{ dBm}$$

$F_w = 2^*F_{tx} + F_{rx}$

When the blocker frequency is between 5950 and 6130 MHz, the CW blocker has a level of –15 dBm. As the duplexer attenuation at the blocker frequency is around 30 dB, P_{cw} is –45 dBm. With P_{im3} = –117 dBm, which is 16 dB below noise floor N_t, IIP3 is calculated as:

$$IIP3 = 0.5(P_{CW} + 2P_{TX} - P_{im3}) = 5 \text{ dBm}$$

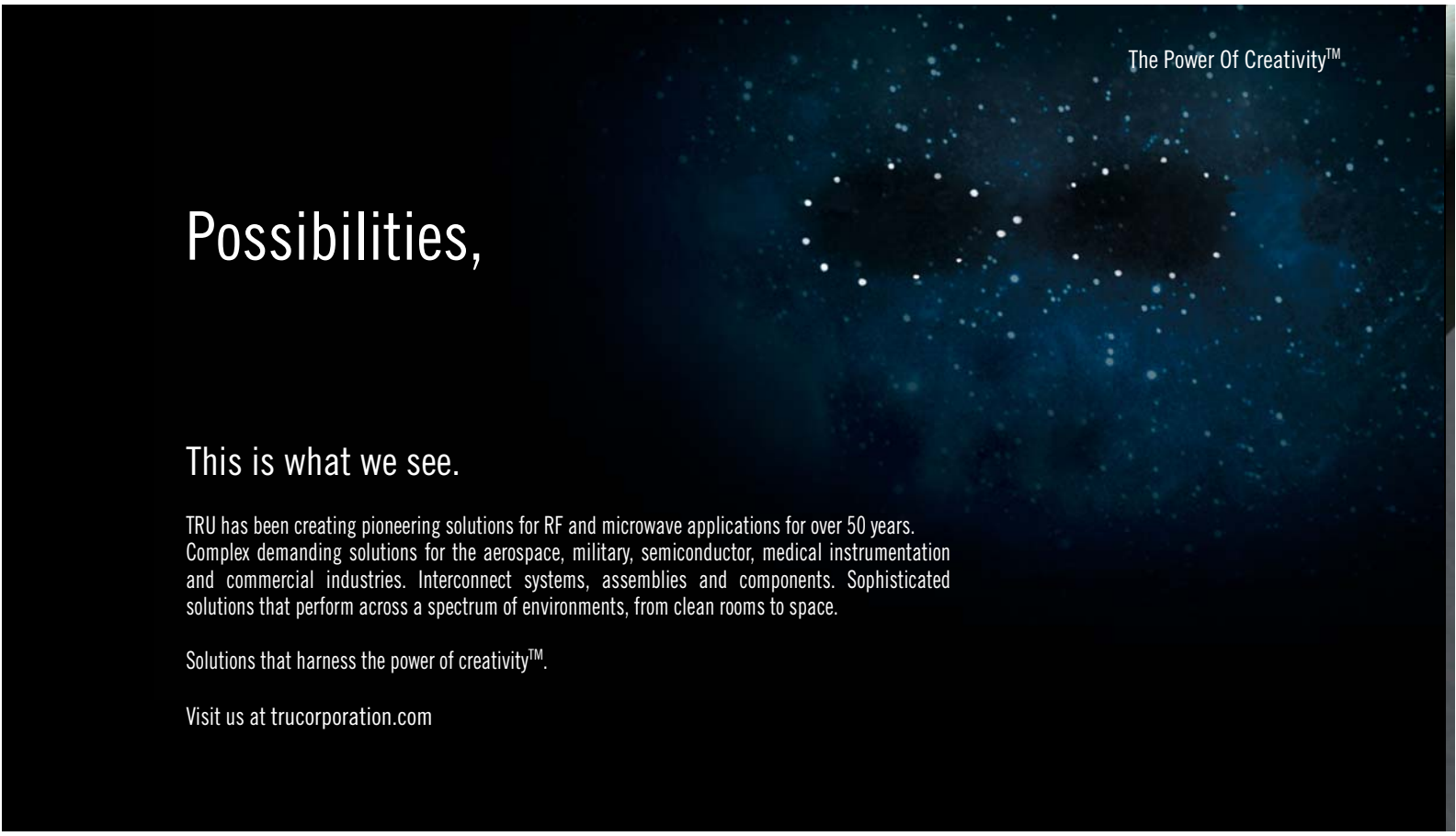
$F_w = \Delta F/2$

In this case the blocker is defined as –15 dBm CW signal, and the duplexer loss at this frequency is 30 dBm. Then P_{cw} is –45 dBm. Similar to the previous case, to have P_{im3} 16 dB below the maximum allowed noise floor N_t, IIP3 is calculated as:

$$IIP3 = 0.5(2 \cdot P_{CW} + P_{TX} - P_{IM3}) = 5 \text{ dBm}$$

V. Measured Results

The IIP2 and IIP3 have been fully verified on a Broadcom WCDMA transceiver IC. The results of IIP2 at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TX frequency and IIP3 wideband intermodulation are shown in Figures 6 and 7, respectively. Both results show that the receiver has excellent linearity.

VI. Conclusion

The system analysis of nonlinearity of a WCDMA receiver is discussed. The formulas for IP2 and IP3 for all test cases are presented and applied. The formulas are useful for system analysis. The assumed margins can be different in each test case. Also, the formulas may be slightly different from other published results due to differences in simulation setup.

Acknowledgment

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Author Information

Chris W. Liu received his M.Sc from University of Montreal, Canada and B.Sc from Tianjin University, China. Chris has more than 17 years experience in the wireless communications. He is currently with Broadcom as Principal Systems Engineer. Chris can be reached by email at: chrisl@broadcom.com.

Morten Damgaard received his M.Sc. degree from University of Aalborg, Denmark in 1989. He has 20 years of experience doing cell phone RF design, semiconductor RF system design, and radio architecture design for GSM, EDGE, WCDMA, and multi-mode transceivers and radios. Morten currently works with Broadcom Corporation in Irvine, CA as a Senior Manager of Cellular RF System Engineering. He holds 31 U.S. patents.

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Developments in Packaging: RFIC, MMIC, LTCC, SiP, SoC

All electronic products are either getting smaller or becoming more complex in their capabilities. To achieve these twin objectives, circuits are now being packaged in a wide variety of miniaturized form factors. The first step in the process occurred many years ago with the evolution of integrated circuits from dual-inline packages (DIP) to SOIC, SSOP, TSSOP, etc. As surface-mount technology grew in popularity, numerous proprietary packaging options were developed for functional modules—mixers, oscillators, couplers, etc. Many of these modules are still in existence, but have been shrunk to the point where they are no longer "modules" but can be treated as individual SMT components.

Recently, packaging had taken two distinct paths—one for "conventional" building block components, the other for highly complex chips or multiple chip subsystems. There are some common elements, mainly in the board attachment methods: pads, solder bumps or solder balls. Differences are in the number of pins, and the physical enhancements needed for mechanical support in larger packages.

An area of significant development is thermal management. RF/microwave ICs do not have the same advantage of digital ICs regarding scaling of power consumption with size. Thus, smaller die and higher density circuits make it difficult to remove heat form areas where currents are concentrated. To address thermal issues, many recently-developed packages include exposed metal pads that provide heat dissipation via the PC board or an external heat sink. A major development is flip-chip technology, where the IC die is mounted upside-down. The substrate is attached to a heat spreader on the top of the package, which has lower thermal resistance than previous packaging methods. Also, the metal layer connections are much closer to the package's external connections, improving high frequency performance.

Advanced research work includes thermal enhancement using carbon nanotubes. An array of these structures can be used to conduct heat from the die, or from specific areas of the die, to a heat dissipating pad on the top or bottom of a package. With the short lengths involved, and encapsulation with a package, this is considered one of the most practical near-term applications for carbon nanotube technology.

To illustrate some of the recent developments in the packaging, we include the following news items:

Package	Description
SOIC	Small Outline IC – 0.05 in. pitch
SSOP	Shrink Small Outline Package – 0.025 in. pitch
TSSOP	Thin SSOP – reduced vertical dimension
QFP	Quad Flat Pack – terminal pins four sides, pitch may be 0.4 mm to 1.0 mm
QFN	Quad Flat No leads – 0.5 mm pitch typical
LCC	Leadless Chip Carrier – usually high pin count, 0.7 mm pitch
BCC	Bump Chip Carrier – Solder bump version of LCC
BGA	Ball Grid Array – very high pin count package, 45 µm pitch typical.
LTCC	Low Temperature Co-fired Ceramic, accommodates a wide range of structure options
Flip-Chip	Die mounted substrate up, metallized top layer down — shorter distance from top-side connections to bumps or solder balls. Substrate may have top-side heat sink.

Many specifications include pin count, e.g. SO-8, TSSOP-24, QFN-24, etc.

Many styles include sub-group or special material notations, e.g., TQFP (thin QFP), PLCC (plastic LCC), LBGA (low-profile BGA), etc.

Package choices are based on die size, operating frequency/speed, pin count requirements, plus size and cost considerations, including compatibility with existing assembly methods. Many IC products are offered with multiple package options.

Table 1 · A partial list of major packaging types for RFICs, MMICs, multi-chip or system-on-chip devices.

LTCC Software to Speed Design

DuPont Microcircuit Materials (http://mcm.dupont.com), part of DuPont Electronic Technologies, and CAD Design Software have announced the integration of DuPont™ GreenTape™ low temperature co-fired ceramic (LTCC) materials and manufacturing processes into CAD Design Software's Electronic Design Automation (EDA) design tools for Ceramic (Hybrid/

Value Added and Cable Assemblies

Now you can rely on Delta, a source of high-quality RF connectors for over fifty years, for all of your cable assembly needs, as well as other connector-related value-added products:

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- One-piece body designs to support attenuator housing needs.
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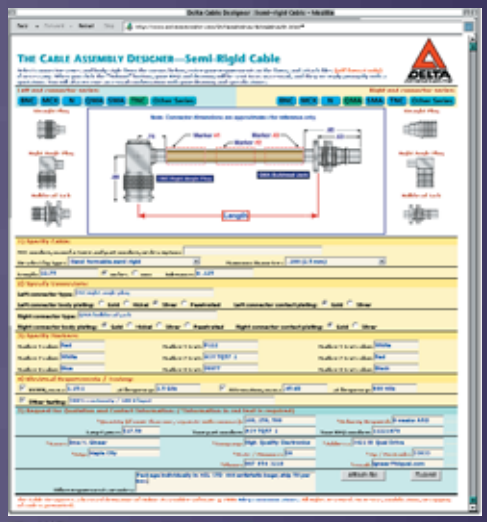
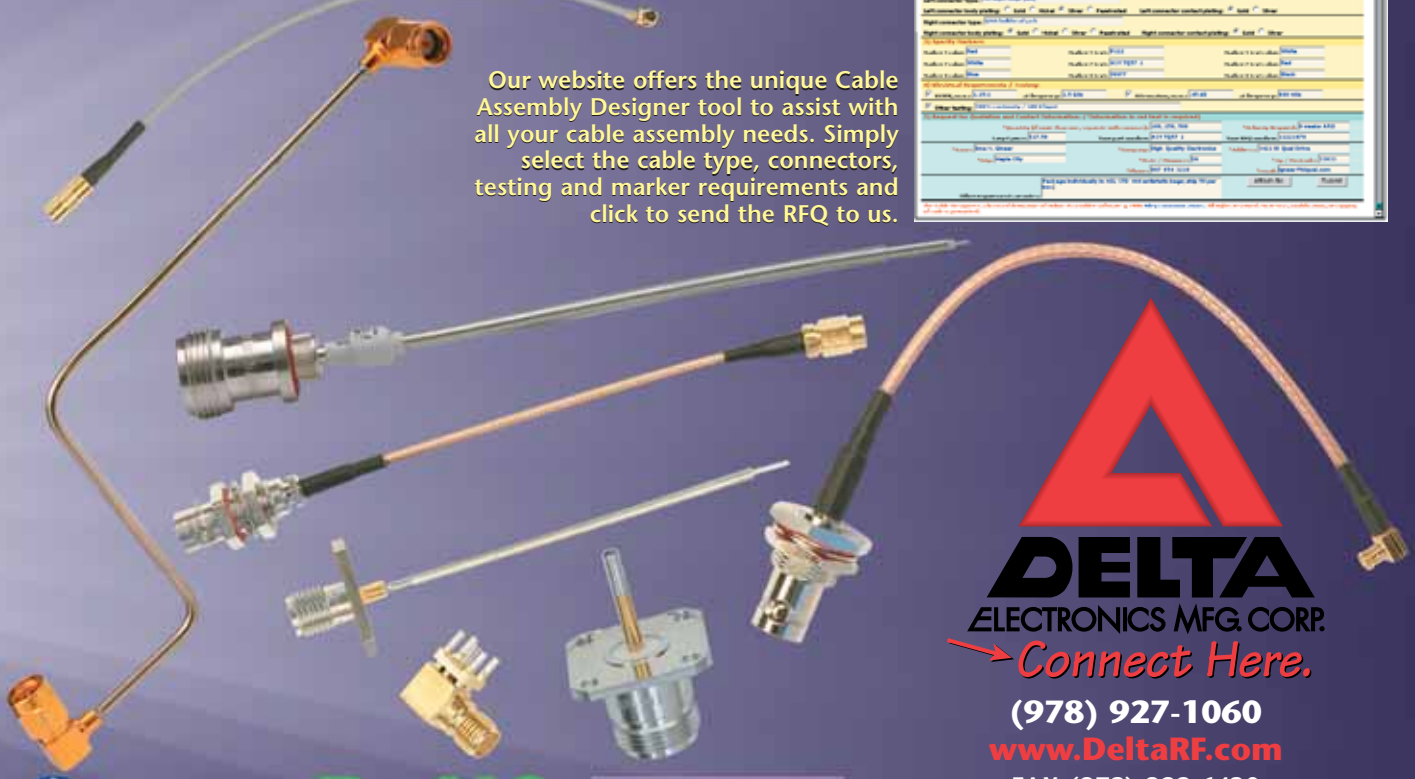


Our website offers the unique Cable Assembly Designer tool to assist with all your cable assembly needs. Simply select the cable type, connectors, testing and marker requirements and click to send the RFQ to us.



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MCM – LTCC) circuit design. By incorporating the widely used DuPont™ Green Tape™ 951 and 943 systems, CAD Design Software's Ceramic Design tool automates DuPont's recommended LTCC processes, and significantly improves cycle time for advanced circuit design.

The ability to select materials from CAD Design Software's Materials Library allows the user to setup custom technology parameters to begin a ceramic design or choose one from the list of preset technology files. The preset technology files have all of the DuPont recommended minimum entity widths and spacings incorporated for efficient design and manufacturing flow. CAD Design Software has incorporated many CAM tools for the ease of processing a ceramic design, such as nibbler data for removing cavity material in bulk or in steps, and control of the Gerber, hole and punch data for correct output to manufacturing.

DuPont Microcircuit Materials has over 40 years of experience in the development, manufacture, sale, and support of specialized thick film compositions.

Fine Pitch Bump Adapters for Alternate ICs

Aries Electronics (www.arieselec.com) has launched a new series of adapters that enable the use of virtually any SMT IC device on a pitch of 0.4 mm or higher on PC boards with 0.5 mm pitch. These are ideally suited to adapting an IC device to some of the more popular TSSOP and QFP packages on 0.5 mm pitch. This adapter will enable the user to solder a BGA or other SMT device to pads on the component side of the adapter, which will then connect through the adapter to the 0.5 mm pitch raised connection pads (up to 0.010") on the bottom. The connection scheme comes standard in a pin 1-to-pin 1 routing, but can easily be customized to accommodate virtually any connection requirement.

The Aries Fine Pitch Bump Adapter boards are fabricated from 0.032" thick FR4 or Rogers 370 HR, with 1/2 oz. copper traces on both sides. The NSMD pads are finished with ENIG. The adapters operate at up to 221°F (105°C) for FR4 versions and 226°F (130°C) for the lead-free version. They are available in a panelized form for easy pick-and-place assembly of the new device, as an adapter only, or as a complete, turn-key solution with devices mounted.

3-D Packaging Initiative

Rohm and Haas Electronic Materials (www.rohmhaas.com), a supplier of materials for semiconductor packaging, has entered into a joint development agreement with IBM to develop and evaluate new materials for emerging packaging technologies. The agreement will focus on the evaluation of photoresists and supporting ancillaries, and low-temperature photodielectric materials for IBM's 3D packaging technologies. Rohm and Haas and IBM will also develop new materials for wafer-level and capillary underfill applications.

The joint collaboration will be performed at IBM's T. J. Watson Research Center in Yorktown Heights, NY, and at Rohm and Haas Electronic Materials' Technology Center in Marlborough, MA.

New Leadframe Technology from DNP

As manufacturers attempt to cut costs, chips are becoming smaller and smaller. Because there is a limit to how short bonding wires can get as chips become smaller, it is necessary to stretch the inner lead to reach the chip. As a result, demand for even finer pitches on the inner lead becomes inevitable. It is believed that the existing minimum lead bonding pitch is 0.17 mm. If chips become any smaller, it is feared that it will become difficult to continue using leadframes, which are very inexpensive for packaging. In response, DNP Electronic Device (www.dnp.co.jp) has developed and begun mass production of a new leadframe, called New-LF, using an innovative process that makes use of DNP half-etching and taping technology.

New-LF is created using selective half-etching on only the fine pitch areas. By doing this, DNP can achieve a fine pitch of less than 0.15 mm without having to make the entire material thinner. The use of this method will enable the future incorporation of 0.18 micron chips and allow for even shorter wires in existing products. As an example, DNP will be able to shorten the wire for each lead on a 60 micron LSI pad pitch, 256-pin QFP to a length of just 4 mm.

Thermally Enhanced Packages

In late 2008, StratEdge (www.stratedge.com) introduced a new family of small outline thermally enhanced molded ceramic packages for power semiconductors. The new line of packages can be used for silicon, silicon carbide, gallium nitride, and other compound semiconductors in power integrated circuit applications. Specific devices include amplifiers, discrete transistors, and diodes where greater than 0.5 watt power is consumed.

These thermally enhanced packages are designed for reliability and to mitigate the inherent stresses of brazing dissimilar materials together. All materials used in the packages have matched coefficients of expansion. They are assembled using a glass-to-metal seal process combined with gold germanium brazing, resulting in a rugged and reliable package. The packages can handle temperatures up to 360° C. A hermetic seal provides enhanced reliability for the device and offers protection from harsh environmental conditions, meeting military standard requirements.

The packages incorporate copper composite bases or copper inserts for enhanced thermal dissipation. Devices are mounted directly to the metal bases thus providing excellent electric ground to the backside of the chip. They provide superior electrical performance for frequencies up to at least 6 GHz. Packages are available in various shapes, sizes, and lead counts. All can be provided with gull wing-shaped leads for surface mounting.

Summary

As this report illustrates, the importance of device packaging means that it is receiving much research and product development attention. Some work is unique to high frequency applications, but there is also much that can be shared with high-volume digital circuits.

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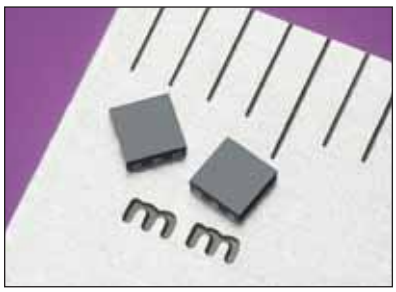

High Gain SiGe:C GPS LNA

In challenging environments GPS receiver ICs with integrated LNAs often come up short when it comes to noise performance and resultant system sensitivity. External LNAs like NEC's new UPC8240T6N provide an effective solution. The UPG8240T6N can be located near the antenna to reduce trace losses and, when combined with distributed filtering, can significantly improve noise performance. Based on NEC's UHS4 (ultra-high speed) SiGe:C bipolar process, the UPC8240T6N is designed to run on supply voltages as low as 1.6 V, so it is ideal for use with new-generation GPS engines. It delivers high gain, low noise and low current consumption. It also features a built-in Power Save function, a robust bandgap regulator, an output matching network, and ESD protection. The UPC8240T6N is housed in a 6-pin, RoHS-compliant 1.5 mm square package.

California Eastern Laboratories
www.cel.com

Network Time Server

Symmetricom, Inc. announced the SyncServer® S300/S350 SAASM NTP Network Time Server that meets the Global Positioning System (GPS) security requirements of the U.S. Department of Defense (DoD). SyncServer S300/S350 SAASM is a high performance time server with a secure, Selective Availability Anti-Spoofing Module (SAASM), GB-GRAM compliant GPS receiver. An ultra high-bandwidth NTP time server, SyncServer S300/S350 SAASM is based on Symmetricom's SyncServer S300/S350 NTP Time Server. The SyncServer S300/S350 SAASM NTP network time servers will be available in July 2009.

Symmetricom, Inc.
www.symmetricom.com

Crystal Oscillators

Bliley Technologies, Inc. has introduced its customizable line of satellite and space qualified crystal oscillators spanning: oven controlled crystal oscillators (OCXOs), temperature compensated crystal oscillators (TCXOs), voltage controlled crystal oscillators (VCXOs), and crystal oscillators (XOs). Bliley's satellite and space qualified crystal oscillators incorporate Z-swept technology meeting the requirements of MIL-PRF-3098. With frequency ranges from 10 to 130 MHz and based on proven designs supplied to military and space customers, Bliley is now manufacturing these devices for multiple customer programs. Key product features include optimized phase noise performance and excellent frequency versus temperature stability.

Bliley Technologies, Inc.
www.bliley.com

Compact Converter/Exciter

Spinnaker Microwave introduced the Model SMS-ACX RF converter/exciter, a compact module that generates signals with modulating bandwidths up to 1 GHz with various combinations of phase, frequency, and amplitude characteristics, and produces an output at any frequency from 10 MHz to 14 GHz, and optionally to 40 GHz. The SMS-ACX is a cost-effective alternative to expensive arbitrary waveform generators (AWGs), and an excellent choice for applications ranging from frequency-agile threat emitters in an EW training platforms, to ATE systems and synthetic aperture radar systems.

Spinnaker Microwave, Inc.
www.spinnakermicrowave.com

Avionics Test Set

Aeroflex announced its IFF-45TS avionics test set has received official certification from the U.S. Department of Defense (DoD) International AIMS Program Office. The IFF-45TS test set is now AIMS certified to perform testing and validation on identification, friend or foe (IFF) Mode 4 and 5 transponders and interrogators. In addition, the IFF-45TS will perform testing on DME/TACAN interrogators. The IFF-45TS test set is an RF signal simulator that can be remotely operated. It supports engineering development of Mode 5 or Mark XIIA systems in bench and over-the-air applications. With appropriate antennas, the IFF-45TS has a test range to 3 km and is capable of ramp testing installed equipment. Orders are being accepted for the now AIMS-certified IFF-45TS test system.

Aeroflex, Inc.
www.aeroflex.com

1250-W Ku-Band Amplifier

Comtech Xicom Technology introduces a new lightweight, outdoor, traveling wave tube amplifier (TWTA). It is the SATCOM industry's first outdoor packaged 1250

watt TWTA designed to operate in the Ku-band uplink frequency. This new amplifier, Model XTD-1250KL, uses a high-efficiency, dual-stage depressed collector, 1.25 kW peak power, TWT to provide 450 W of linear power at the flange, over the full 13.75 to 14.5 GHz uplink frequency band.

Xicom Technology, Inc.
www.xicomtech.com

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Agilent Technologies, Inc.
www.agilent.com/find/mimo-forward

LTE Test Solutions

Aeroflex announced it is on track to become the a complete global supplier of LTE test equipment by bringing to market solutions covering all aspects of LTE test. Aeroflex has demonstrated its ability to deliver early solutions for the design, development and deployment of both LTE terminals and network infrastructure with the Aeroflex 7100 and TM500 respectively. Aeroflex has also LTE-enabled its 3000 Series PXI-based and 3410 digital RF signal generators. The Aeroflex 7100 delivers the most comprehensive testing capability for LTE mobile devices available in a single bench-top instrument enabling chip-set designers, software developers and handset manufacturers to accelerate the pace of development projects. The Aeroflex TM500 has become the industry's de-facto test mobile fully supporting LTE network development, verification and optimization, both in the lab and over the air. Supporting LTE-FDD and TD-LTE standards, the Aeroflex TM500 provides single, multiple and capacity test mobile versions for testing against LTE base stations and network.

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IF/RF MICROWAVE COMPONENTS

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3-D EM Simulator is Integrated with ADS to Lower the Cost of Design

By How-Siang Yap & Hee-Soo Lee
Agilent EEsof EDA

With EMPro 2009, high performance time domain (FDTD) and frequency domain (FEM) EM analysis is integrated with all the other Advanced Design System (ADS) capabilities

Full 3-D electromagnetic (EM) simulation used to be an analysis tool for EM experts tackling the toughest antenna or passive component design problems. Today, however, consumer wireless technology trends are pushing high-frequency and high-speed circuits, packages, modules, and antennas into ever smaller form factors. The resultant 3-D EM proximity effects now must also be considered by circuit and module designers. Discovering and fixing these proximity issues during hardware testing often occurs very late in the design process, which is too expensive.

EMPro 2009, the new EM platform from Agilent EEsof EDA, brings powerful 3-D EM simulators within an efficient and intuitive 3-D design environment that is easily accessed by circuit and module designers in the industry-leading Advanced Design System (ADS) design flow. The dedicated 3-D design environment enables the intuitive creation of 3-D parameterized components such as metal shields, packages, laminates, dielectric bricks and interconnects for insertion into the ADS 2-D circuit layout environment. The resulting combined circuit-3-D EM simulation will then reveal any unexpected interactions among the components. Unlike a design process that uses standalone 3-D EM tools, the parameterized 3-D EM components in ADS enable quick and convenient adjustments to the component geometries, providing insight into the EM effects on the design without leaving the design flow.

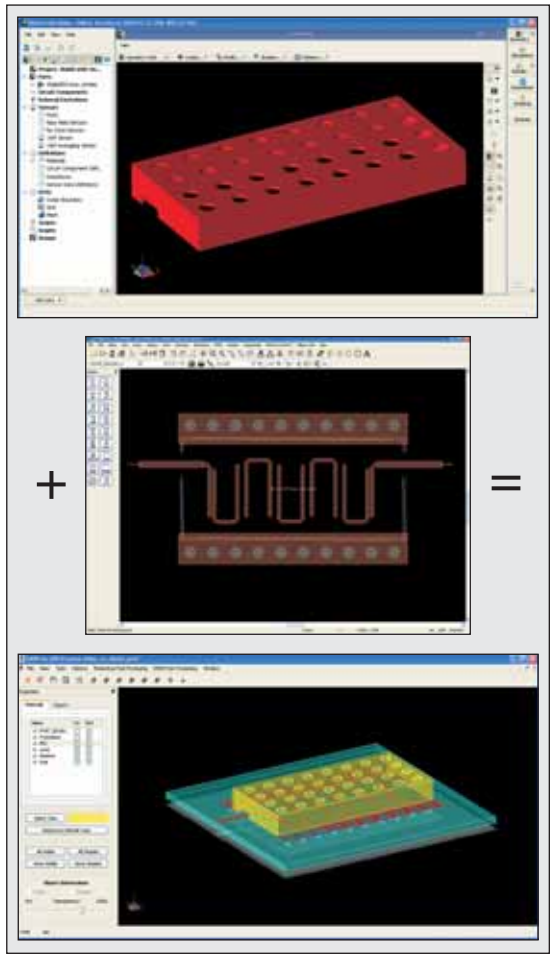

Figure 1 · EMPro 2009 creates a 3-D parameterized component (metal shield) that integrates with the ADS circuit (a filter) for fast, efficient simulation and optimization.

In the past, using point tools, designers needed to leave the design flow to perform multiple 3-D drawings, design setups, re-sim-

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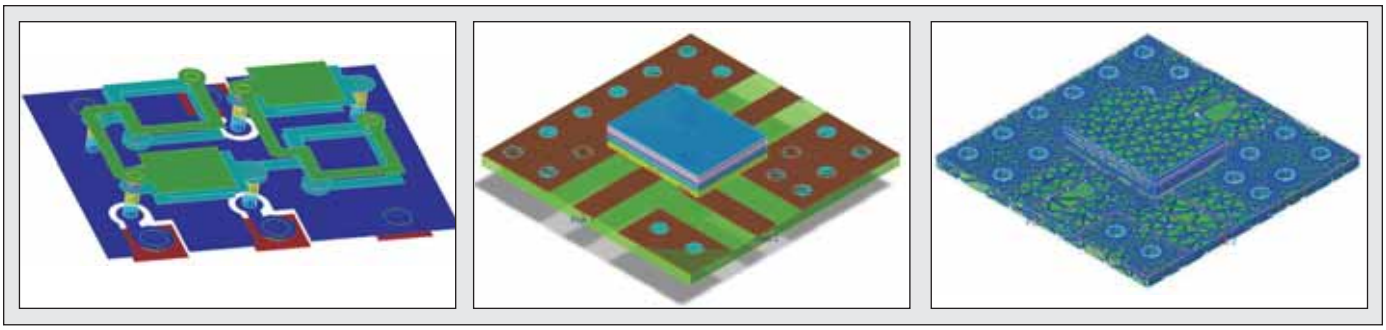

Figure 2 · An example LTCC Balun, showing the module, the LTCC Balun mounted on test board, and the Balun and test board meshed for full FEM 3-DEM simulation.

ulations, and error-prone manual imports. EMPro's 3-D EM integration into the circuit or module design flow can save designers hours per simulation. This can dramatically lower the overall cost of design.

The 3-D EM design environment in EMPro 2009 is designed to provide the fastest route from 3-D drawing to accurate simulated results. The key capabilities include:

- Two powerful 3-D EM simulators in one platform: Finite Element Method (FEM) and Finite Difference Time Domain (FDTD) simulators cover the broadest range of applications, from electrically small components to electrically large antennas. They also provide convenient cross-validation from both time- and frequency-domain 3-D EM analyses for the same structure, helping designers to gain insight and to have greater confidence in their designs without needing to laboriously re-enter designs and simulation setups in separate tools.
- Efficient import, parameterization, and setup of complex 3-D CAD designs for 3-D EM simulation. For example, multiple simulations over a family of geometries for a complete cell phone can be done in one batch, without long repetitive manual setups for each simulation.
- "Set-it-and-forget-it" material assignment. Drag and drop assignments onto the imported 3-D object from an extensive material database to set up the simulation. Subsequent CAD imports inherit the assigned material properties for even faster simulation startup.
- Interactive 3-D editing with intuitive context-sensitive on-screen cues remove the constant need to look away from the drawing to select from cluttered menu picks and dialog boxes, allowing designers to complete the 3-D drawing quickly and efficiently.
- Direct parameterization by simply typing over any numerical dimension with an equation variable, the value of which can further depend on multiple other variables. This enables linked geometry sweeps or drawing mathematically defined surfaces for antennas to obtain the optimal performing geometry.
- Adaptive meshing follows the shape of the critical components to always deliver the best accuracy and convergence in a 3-D EM simulation. It remains active during parameterized sweeps to ensure the most accurate simulation possible.
- Dynamic real-time plots during simulation allows designers to begin see EM effects as the simulation runs and without waiting for it to complete. Designers can interrupt the simulation to adjust the geometry and the built-in intelligent simulation setup will restart with minimal user input.
- Intuitive setup of EM field sensor locations and immediate viewing of results for antenna far field, Specific Absorption Rate (SAR) and Hearing Aid Compatibility (HAC) through multi-threaded accelerated post processing, eliminating the long wait in traditional 3-D EM simulators, where post processing can take as long as the simulation itself.

EMPro Application Examples

The following examples of typical RF modules, interconnects, antenna designs, and packages demonstrate the value of EMPro's capabilities for achieving a faster, more accurate engineering design flow in the ADS environment.

LTCC (Low Temperature Co-fired Ceramic) or Laminate RF Modules

RF modules typically are constructed from multilayer ceramic or laminate dielectric material with embedded RF passive components between the layers. Such dielectric brick structures cannot be accurately solved by planar 3-D EM simulators, which assume infinite dielectric layers and do not account for edge proximity fringing. The embedded RF components are drawn by RF circuit layout macros which would be very time consuming to reproduce in a standalone 3-D EM tool. Hence FEM full 3-D EM simulation integrated within the circuit design flow is the ideal solution for these applications.

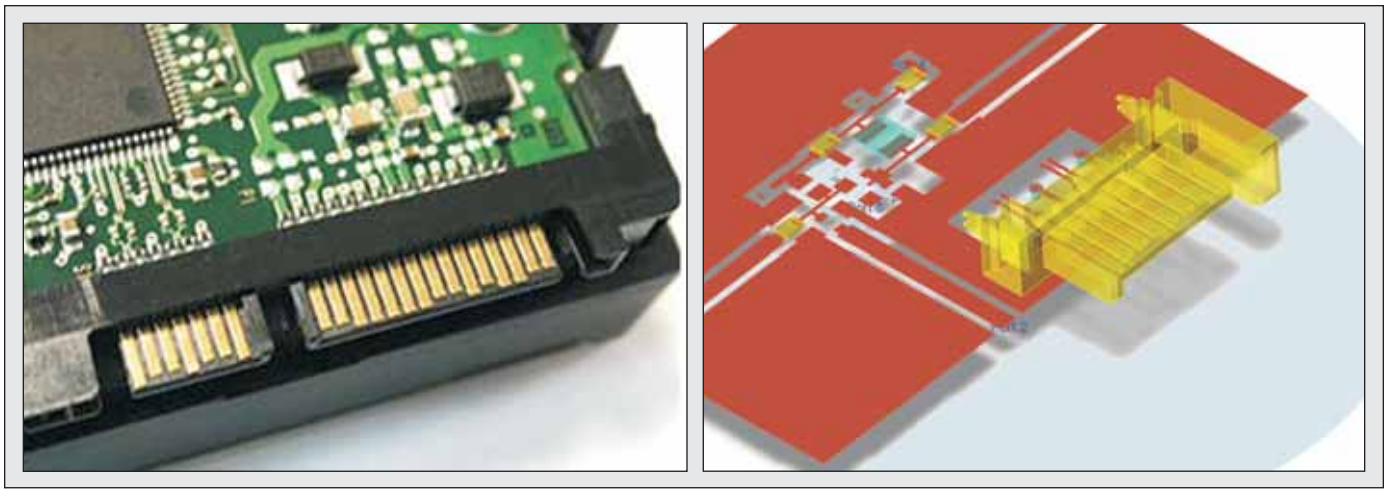

Figure 3 · SATA connector modeling in EmPro.

High-Speed/High-Frequency Connectors

High-speed and high-frequency interconnects are now an integral part of the digital interface designs for PCs, peripherals, and portable computing devices such as netbooks and smart phones. The following example (Figure 3) shows the Serial Advanced Technology Attachment (SATA) Connector in disk drives that potentially has to support 6 Gbits/s data throughput. The *S*-parameter models of the SATA connector family are developed in EMPro and can be cross-verified with the included FEM and FDTD simulators to give designers double confidence in 3-D EM simulation accuracy. The models are contained in an ADS design kit that can be distributed and installed into ADS

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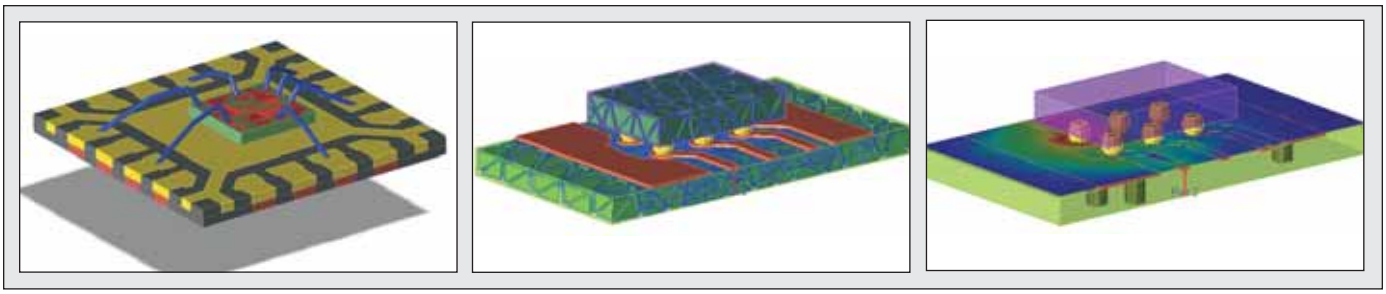

Figure 5 · Analysis of the 3-D EM effects of bond wires, flip-chip solder balls/bumps and packages can be included in the MMIC/RFIC and System-In-Package (SIP) design flow.

as a connector library for use in signal integrity analysis and design of high-speed serial channels.

The next example shows a simulation of the antenna performance in wireless consumer products such as 4G smart phones, netbooks, and routers under realistic operating surroundings that include the product housing, battery pack, PCB boards, and proximity orientation relative to the human hand and head. It is quite remarkable that such a simulation is even possible, but with GPU accelerated FDTD simulation and multi-threaded post processing, results like those shown in Figure 4 can be obtained in approximately one hour.

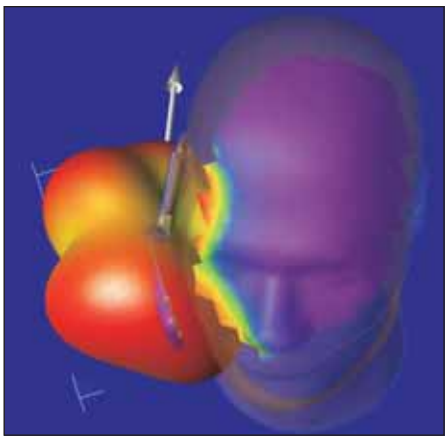

Figure 4 · EMPro allows designers to simulate antenna performance under real world conditions in a cell phone enclosure including human head and hand proximity to determine SAR, HAC (hearing aid compatibility), MIMO (multiple input, multiple output) correlation and feed impedance.

With these fast computation times, it is an easy matter to examine multiple orientations of the antenna, representing different human user handling positions, providing the necessary data for the design of adaptive antenna matching circuits in ADS. The temperature rise in the human head due to cell phone antenna radiation can also be checked for compliance against industry safety Specific Absorption Rate (SAR) standards, so that expensive rework on the actual phone during hardware testing can be avoided.

IC Interconnects to Package, System-In-Package with Wire Bonds, or Flip-Chip Solder Ball

The design of an RFIC, MMIC, or SIP is not complete until the effects of packaging and the interconnects between them such as wire bonds, solder balls, or solder bumps are considered. Traditionally, designers had to draw and analyze them in a separate, 3-D EM tool and then laboriously import the results back to the IC or SIP circuit design environment for a combined analysis. Now, you can efficiently create these 3-D components in EMPro to be used with 2D circuit layouts in ADS in the co-design of IC, package, laminate, and module with circuit and 3-D EM simulation in a streamlined design flow.

Summary

Unlike standalone FEM or FDTD 3-D EM tools, EMPro 2009 includes both FEM and FDTD simulators under the industry's most modern and efficient 3-D EM design environment to cover the widest range of applications. It allows for cross-verification of time- and frequency-domain 3-D EM results to double the designer's confidence in the simulation.

Its integration with ADS, the leading high-frequency circuit design platform, provides a combined 3-D EM and circuit design flow in a single high performance design system, at substantially lower cost than that of owning and maintaining a collection of standalone tools.

Author Information

How-Siang Yap is the product manager for the Advanced Design System EDA platform in Agilent EEsof. His team develops design solutions for the RF-SiP, MMIC, RF-mixed signal board, antennas and RF/microwave components targeted at the commercial wireless and defense-aerospace industries. He can be reached by email at: how-siang_yap@agilent.com.

HeeSoo LEE is 3D EM (Electro-Magnetic) simulation technical lead at Agilent Technologies EEsof EDA. In his 20-year career, he has worked for RF/microwave EDA companies including Agilent EEsof, Cadence and Hewlett-Packard.

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Linearity Improvement Techniques for Wireless Transmitters: Part 2

By Andrei Grebennikov
Bell Labs Ireland

This two-part article presents a wide range of techniques for amplifier linearization, along with historical notes to help us appreciate the creative work in their development

Reflect Forward Linearization Amplifier

The reflect forward adaptive linearizer (RFAL) technique uses the input reflected signal from one power amplifier to develop an intermodulation-correcting signal forwarded to the input of the other power amplifier which are connected in parallel or balanced configuration [23, 24]. Figure 6 shows a block diagram of the RFAL amplifier which includes the two identical power amplifiers, a low-distortion booster amplifier, forward and reflected paths with two directional couplers at the input, and delay line and in-phase combiner at the output. With the RFAL technique, the power combining efficiency of the two power amplifiers approaches that of the conventional parallel power amplifier configuration, while the intermodulation products can be improved from 20 to 30 dBc at the center bandwidth frequency.

The basic principle of RFAL operation can be described as follows: when the two-tone forward fundamental signal is reflected from the transistor input of upper power amplifier, the resulting reflected two-tone fundamental signal is out-of-phase relative to the input forward signal and is in-phase relative to the output signal. In this case, the intermodulation components appeared at the upper power amplifier input as a result of its active device nonlinearity are in-phase relative to the output intermodulation components. The input reflected composite signal containing the fundamental and intermodulation components is used as a correcting signal for the lower power amplifier. This correcting signal, when properly amplified and phased in the reflected path, cancels the output intermodulation distortions produced by the lower power amplifier when it is combined with the input signal flowing through the forward path. The booster amplifier in a lower path is necessary to equalize the drive signal levels for both upper and lower power amplifiers. In this case, the signals from the upper and lower amplifying paths have the in-phase fundamental and out-of-phase intermodulation components at the corresponding inputs of the output in-phase combiner, thus resulting in a distortion cancellation in the combined signal flowing into the load. The LDMOS RFAL amplifier with output power of 43 dBm achieves an improvement of the third-order intermodulation products by over 15 dB and a total efficiency of more than 20% over the frequency range from

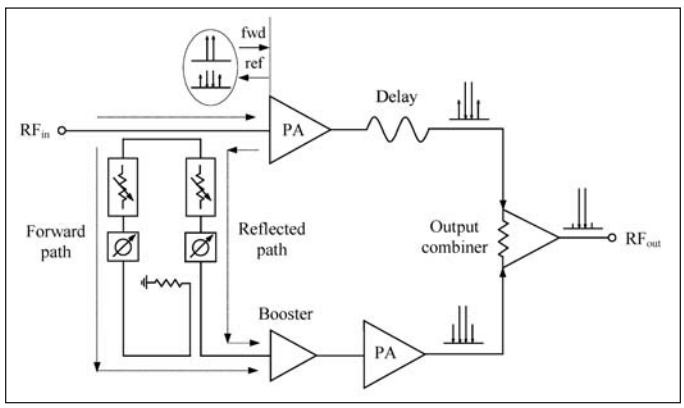

Figure 6 · Block schematic of reflect forward linearization amplifier.

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865 to 895 MHz [25]. For very high cancellation requirements over wide temperature and drive conditions an adaptive feedback network can be used to monitor the relative amplitude and phase at the outputs of the power amplifiers and drive the voltage variable attenuators and phase shifters to maintain the optimum conditions.

Predistortion Linearization

To achieve simultaneous high-efficiency and low-distortion operating conditions of the power amplifier when the linearity requirements are not extremely high, it is possible to use a predistortion linearizer, which provides the positive amplitude and negative phase deviations for input RF signal to compensate for the active device nonlinearity. This is possible since nonlinear behavior (when a power amplifier is operated close to saturation) usually represents the opposite behavior of its amplitude and phase characteristics. Historically, the initial idea to compensate for the third-order intermodulation products arising in vacuum-tube amplifier was to use the linearization scheme where a nonlinear amplifier having a compressing characteristic is followed by a nonlinear element having an expanding characteristic and producing the third-order distortion of opposite sign to that of the amplifier [26]. The block diagram of the linearized power amplifier system with a predistortion linearizer with indication of the appropriate amplitude and phase dependences at each stage of the system is shown in Figure 7, where a variable attenuator for adjusting the amplitude level of the input signal is also included. At microwaves, a linearized power amplifier usually includes two isolators for stable operation conditions. The conventional predistortion linearizer circuits generally use either diodes or transistors as sources of intermodulation [27, 28].

As an interesting fact, as early as the beginning of the1920s, it was claimed that, by using similar vacuum tubes in both stages of a two-stage power amplifier with similar signals at their inputs, the even harmonics generated by the first amplification are neutralized by the even harmonics generated by the second amplification because they are similar in amplitude and opposite in phase at the output of the second vacuum tube [29]. Indeed, as it turned out with regard to modern transistor power amplifiers using GaAs pHEMT devices, it is enough to choose a proper bias point for a driver-stage device in a two-stage amplifier to provide a negative phase deviation to compensate for the positive phase deviation of the final stage [30]. In this case, the quiescent current of the driver-stage device, whose size is three times smaller than that of a final-stage device, is sufficiently small. As a result, for a quiescent current equal to 1.25% of the device DC saturated current, an improvement of more than 5 dB in *ACLR* of a whole high-efficiency two-stage cellular-phone WCDMA power amplifier operating at 1.95 GHz can be achieved at backoff output powers close to the saturation power.

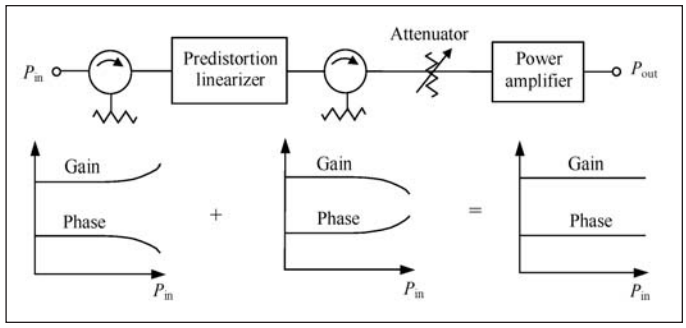

Figure 7 · Block diagram of power amplifier with predistortion linearizer.

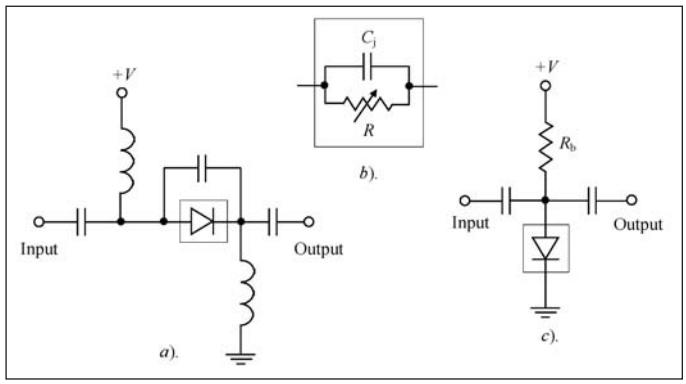

Figure 8 · Simple diode-based predistortion linearizers.

Figure 8(a) shows the schematic of a simple diode linearizer composed of a series Schottky diode and a parallel capacitor with two RF chokes for DC feed and two blocking capacitors, which provides positive amplitude and negative phase deviations when input power increases [31]. The equivalent circuit of the series diode is shown in Figure 8(b), where R is the diode equivalent resistance and C_j is the junction capacitance. With the increase of an incident input signal power, the forward diode current increases that leads to the decrease of the diode resistance R. In this case, the positive amplitude and negative phase deviations can be achieved under low forward-bias conditions when the diode current ranges from 0.1 to 1.0 mA, and, in the latter case, the phase deviation can reach a value of –30°. Applying such a linearizer to a 1.9 GHz MMIC power amplifier with saturated power of 22.5 dBm, an improvement of adjacent channel power ratio (*ACPR*) of 5 dB can be achieved for the QPSK modulated signal when output powers are less than 15 dBm.

A similar improvement of *ACPR* can be achieved by using a linearizer based on a parallel Schottky diode with the bias feed resistor R_b, which is shown in Figure 8(c) [32]. With the increase of input power, the bias point of a diode changes due to the voltage drop across the resistor R_b caused in turn by the increased diode forward current.

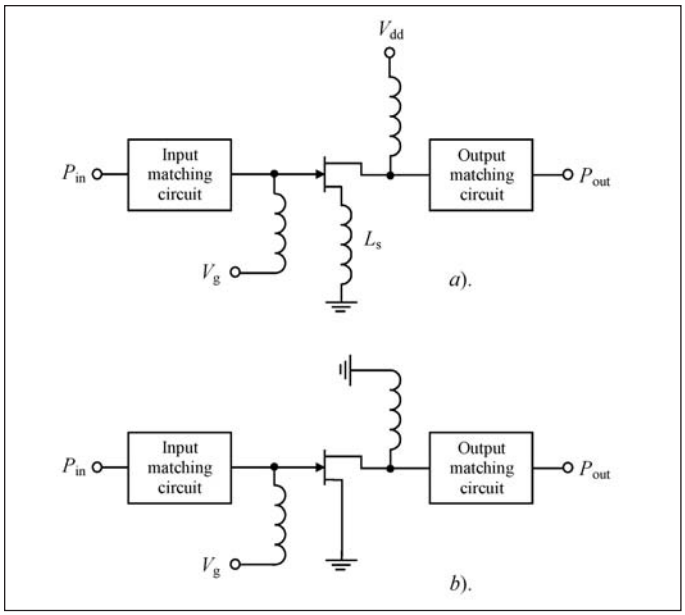

Figure 9 · Transistor-based linearizers.

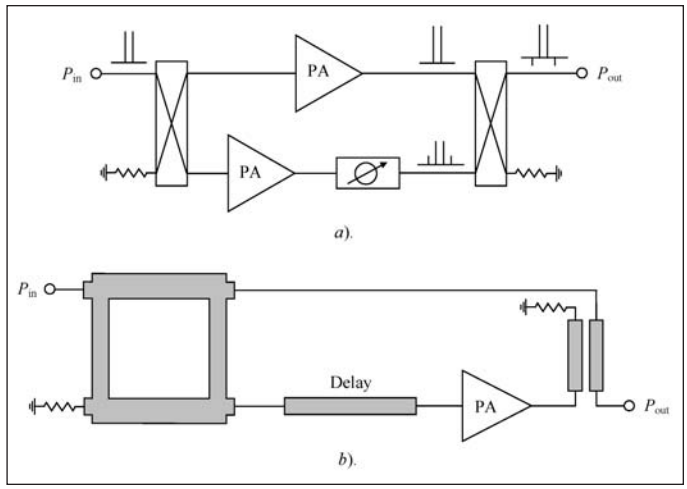

Figure 10 · Block diagrams of power amplifier linearizers with input power splitting.

As a result, due to the decreased diode resistance R, the linearizer achieves positive gain and negative phase deviations. By applying such a linearizer to a 2.7 GHz power amplifier, a maximum improvement of 5 dB was achieved for low quiescent current conditions at output power of 34 dBm.

Positive amplitude deviation with negative phase deviation can also be achieved using a series-feedback GaAs MESFET amplifier with a large source inductance L_s, a block diagram of which (including matching circuits) is shown in Figure 9(a) [33]. The required amplitude and phase deviations are due to nonlinearities of the device transconductance gm, gate-source capacitance C_{gs}, and differential drain-source resistance R_{ds}. For the device with a gate width of 1.2 mm, a nonlinearity of g_m contributes to the positive amplitude deviation when L_s = 20 nH. At the same time, nonlinearities of both g_m and R_{ds} contribute to the negative phase deviation when $L_s \geq 3$ nH. A nonlinearity of C_{gs} has a negligibly small effect on both the amplitude and phase deviations. As a result, for a linearizer with L_s = 16 nH at an operating frequency of 1.9 GHz, the positive amplitude and negative phase deviations were obtained across the input power dynamic range from 5 to 18 dBm, with amplitude deviation of 2.5 dB and phase deviation of 30° at 18 dBm input power. The GaAs MESFET device was biased in Class AB mode with a drain-source supply voltage of 2 V providing a quiescent current of 78 mA. By applying this linearizing technique to a 1.9 GHz MMIC power amplifier with 1-dB compressed power of 17 dBm, an improvement of *ACPR* up to 7 dB was achieved for a π/4-shifted QPSK signal.

As an alternative, it is also possible to achieve positive amplitude and negative phase deviations using a source-grounded MESFET device with zero drain-source supply voltage [34]. The schematic diagram of such a linearizer is shown in Figure 9(b). In this case, for the device with a gate width of 240 mm at the saturation power of 20 mW under the gate bias condition of V_g = –0.4 V, the 3-dB increased power gain and of about 30-degree negative phase were achieved using the varying drain-source resistance. Because of its simplicity, such a linearizer can operate from 2 to 12 GHz with good thermal stability. When it was implemented into a 50 W solid-state power amplifier system at operating frequency of 7 GHz, the system noise power ratio was improved over 15-dB dynamic range, in particular by 2 dB at the 3-dB output power backoff point.

A more advanced configuration of the predistortion linearizer is based on the splitting of the input signal into nonlinear and linear paths using a directional coupler or a hybrid divider with subsequent subtraction of the resulting signals in the output coupler-subtracter. The block diagram of such a predistortion linearizer which employs two power amplifiers in a balanced configuration using two 90-degree hybrids is shown in Figure 10(a) [28]. In this case, the upper power amplifier is operated in a linear Class A mode while the lower power amplifier is biased in a nonlinear Class AB or B mode to generate the proper intermodulation products by controlling the input power and device bias conditions. The phase shifter in a lower amplifying path is necessary to optimize the level of the fundamental components in the resulting output spectrum. Since both devices present approximately the same input impedances, a low input return loss is provided because the most of the reflected power flows into the isolated port. Figure 10(b) shows the practical microwave microstrip implementation of a two-path predistortion linearizer with an input 90-degree branch-line hybrid

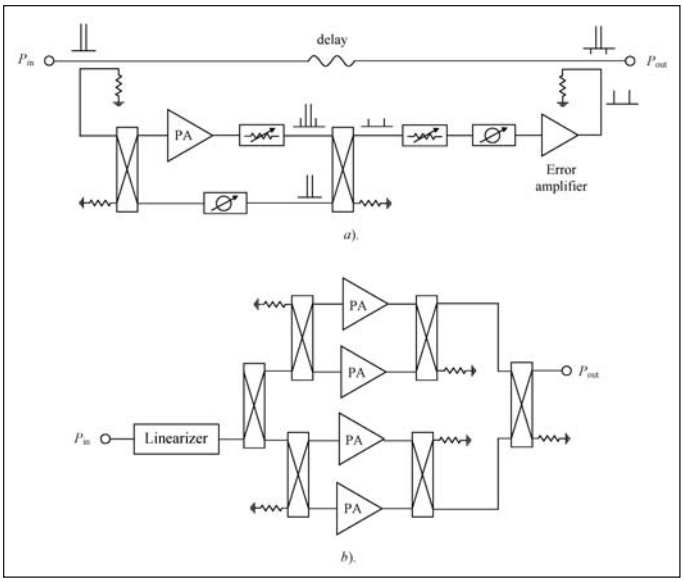

Figure 11 · Power amplifier module with linearizer.

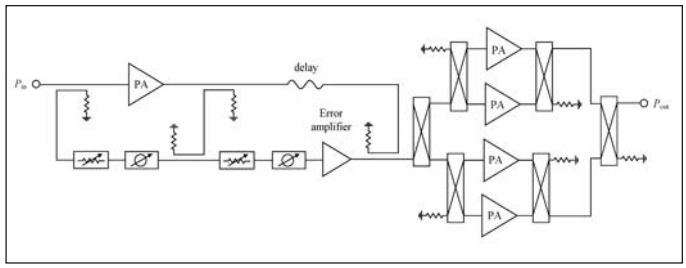

Figure 12 · Power amplifier module with feedforward predistortion linearizer.

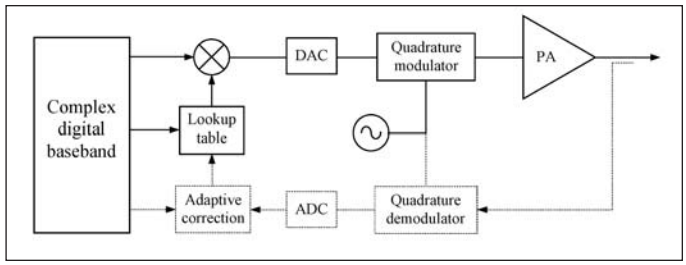

Figure 13 · Digital predistortion system.

coupler, a nonlinear power amplifier in a lower path and an output directional coupler [35]. The microstrip transmission line in a lower amplifying path having a required electrical length compensates for the additional phase shift provided by the active device, whereas the required amplitude conditions are realized with the coupling coefficient of an output coupler-subtracter to be chosen. As a result, for a Ku-band multicarrier 4.5-W power amplifier, the phase deviation of a 12-dBm signal at the linearizer output up to –10° was achieved with a 22-dBm signal at the linearizer input.

Figure 11(a) shows the modified three-path predistortion linearizer structure, where a balanced configuration with a nonlinear power amplifier is adjusted for suppression of the fundamental components with the resulting error signal. Then, the amplitude-adjusted and properly phased error signal is amplified by an error amplifier and added to the linear component in the upper path which is a delayed portion of the input signal. However, it is very difficult to match the nonlinear characteristics of the predistorter and the main power amplifier, because generally they differ both in size and number of stages, which can only result in less than 10 dB improvement of adjacent channel leakage power ratio (*ACLR*) at 5 MHz offset from the center bandwidth frequency [36]. Therefore, it is very important for further linearity improvement to use similar devices in the predistorter and in the main power amplifier with a preferred balanced structure. As an example, the block schematic of a power amplifier module which includes a three-path predistortion linearizer and a main power amplifier based on four power amplifiers configured into a balanced structure is shown in Fig. 11(b). In this case, when the same transistors are used both in the linearizer and main balanced power amplifier operated in Class AB mode, a 12-15 dB improvement of *ACLR* for the signal with peak-to-average ratio (PAR) of 6.5 can be achieved, depending on how close the device operation mode is to saturation.

The concept of a feedforward loop with its high cancellation performance can be used for a predistortion linearizer implementation. Since the feedforward loop is placed in front of the main amplifier, the linearity and power requirements of the error amplifier are reduced significantly compared to the conventional feedforward system. In this case, the delay-line and coupler losses are less significant factors affecting the amplifier performance. Figure 12 shows the simplified schematic diagram of a power amplifier module with the feedforward distortion linearization using five identical power amplifiers based on MRF5S21090 LDMOS devices [37]. For a forward-link four-carrier WCDMA signal at 2.35 GHz, the *ACLR* was enhanced by about 7 dB at 5 MHz offset and the total efficiency of 12.7% was achieved at an average output power of 47.8 dBm, backed-off by 10.8 dB from the total peak power of 720 W.

Figure 13 shows the block schematic of a digital predistortion linearizer where the predistortion algorithm is based on an initially measured PA amplitude-to-amplitude modulation (*AM-AM*) and amplitude-to-phase modulation (*AM-PM*) response extracted from the *S*-parameter measurements by a vector network analyzer (VNA) [38]. The amplitude and phase characteristics are interpolated using splines, which are continuous piecewise cubic functions with continuous first and second derivatives. The

interpolated amplitude and phase characteristics are then used to compute the appropriate predistortion coefficients representing a lookup table which are multiplied with the original IS-95 signal to generate the desired predistorted baseband signal. The results show the limitations of this technique when the LDMOSFET power amplifier operation conditions are close to saturation, only a little more than 6-dB improvement in *ACPR* can be achieved. Generally, an adaptive correction mechanism is required to maintain the performance over varying load, supply voltage, or temperature conditions. This means that the lookup table needs to be updated continuously to keep differences between the source signal and the transmitted signal sufficiently small. This can be realized by downconverting the portion of the transmitted signal and comparing it with source signal. In this case, it is important to provide the optimization of the wordlengths required in different parts of the predistortion linearizer to reduce power consumption and increase bandwidth for the required adjacent channel interference level [39]. The feedback complexity can be reduced with special adaptation algorithm when a single mixer and ADC are used in the feedback path instead of the full quadrature demodulation [40]. In addition, a non-iterative adaptation method can be used to eliminate the convergence constrains usual for iterative methods [41].

Feedback Linearization

The principle of feedback linearization of the power amplifier at the carrier frequency was invented by Harold S. Black in 1927. A year later he filed the patent application on a vacuum-tube feedback amplifier [42]. Black recognized that by using a large amount of feedback in an amplifier comprising several vacuum-tube stages in cascade to yield a very high open-loop gain gives a glorious opportunity to make a negative feedback amplifier having increased bandwidth, and which is insensitive to nonlinearity and uncertainty in the characteristics of the vacuum tube [43]. The gain of the negative feedback amplifier decreases by amount of the feedback or loop gain; so do the nonlinear components. In this case, the negative feedback amplifier becomes insensitive to the gain or phase variations as long as its stability conditions are satisfied. In the latter case, if each of three tuned circuits of a three-stage vacuum-tube RF feedback amplifier can be assumed to have the phase-gain characteristics of the interstage circuits very nearly to infinite Q-tuned circuits, the maximum amount of feedback will allow a phase margin of 30° from a total phase shift of 180° [44]. Unfortunately, the significance of this invention, as well as the operation principle of a negative feedback amplifier, was not fully understood at that time. For instance, Black's director of research insisted that a negative feedback amplifier would never work, similarly the Patent

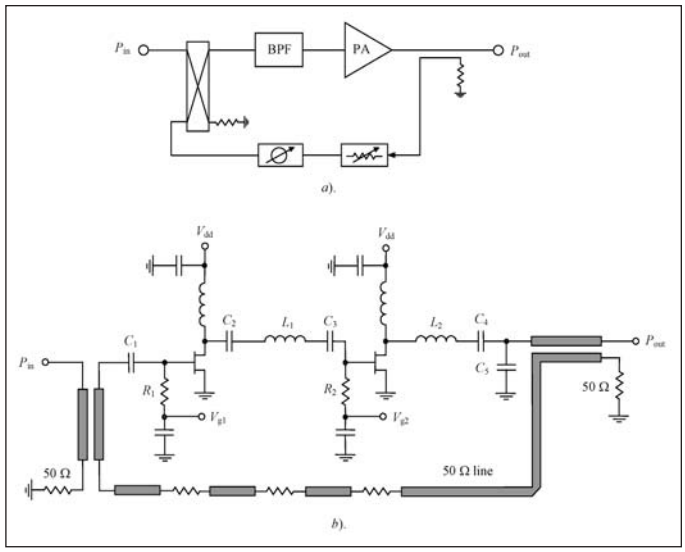

Figure 14 · Negative feedback power amplifier schematic.

Office initially did not believe it would work and took over nine years to decide to issue the patent [45]. Even today, Black's pioneering role for further achievements in feedback theory and practice is not well known.

The basic structure of a negative feedback amplifier at microwave frequencies is shown in Figure 14(a) where the power amplifier, bandpass filter (BPF) and feedback loop elements are chosen to provide a loop transmission greater than one with a phase shift of 180° within the operating bandwidth [46]. In this case, the BPF must be a single tuned resonator, so that the phase shift introduced will be less than 90°. By using a 50-dB gain power amplifier, an improvement of the third-order intermodulation components by 10 dB (from –30 dBc down to –40 dBc) at midband frequency of 2 GHz with the power gain of the closed loop system of about 34 dB and output power of 33 dBm was achieved [47]. It should be mentioned that the feedback power amplifier designed to operate at 835 MHz in communication system can provide a distortion improvement of 10 dB at saturated region, if the feedback gain of at least 20 dB is achieved [48]. Although, a feedback gain of 10 dB results in approximately 10-dB improvement in adjacent channel leakage power at 3 dB power backoff from saturation. In contrast, for the simple resistive feedback single-stage power amplifier, the level of the third-order IMD components at 2 GHz can be reduced by about 5 dB for the medium-signal levels only [49].

Figure 14(b) shows the circuit schematic of the two-stage power amplifier designed for a 3.4-4.2 GHz frequency bandwidth [50]. The output power of 27 dBm was achieved by using a power MESFET device with a gate length of 0.5 mm in the second stage. Loop gain was adjusted by changing the coupling in the output direc-

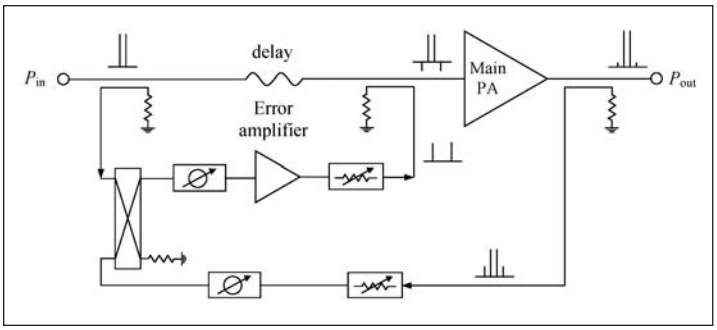

Figure 15 · Power amplifier module with feedback-feedforward linearization.

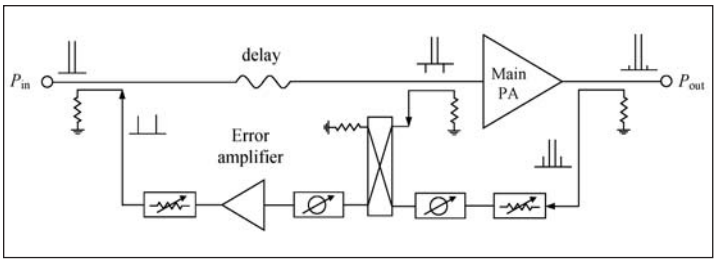

Figure 16 · Power amplifier with feedback predistortion linearization.

tional coupler, whereas a phase shift of 180° in the feedback path at midband was obtained using a microstrip line with the proper electrical length. The predicted open-loop gain was 19 dB at 3.7 GHz, decreasing to 13 dB at the upper bandwidth frequency. The shunt resistor at the input of the first MESFET device is necessary to improve the input return loss. As a result, for the closed-loop configuration, an improvement of the third-order intermodulation components of 7 to 9 dB over a 750 MHz bandwidth at –3 dB power backoff with a power gain of about 10 dB and an input return loss of more than 10 dB.

To improve the gain capability of the negative feedback amplifier, it is possible to combine its simplicity with a feedforward technique to provide a separate path for the error signal only. Figure 15 shows the schematic of the power amplifier module with a feedback-feedforward linearization when the intermodulation distortion products at the main PA input (symbolically plotted as out-of-phase) should be phased in a way that they cancel the intermodulation distortion products at its output. This modified negative feedback circuit can provide a reduction in the intermodulation distortion products equal to the traditional negative feedback topology without the usual reduction in overall amplifier gain [51]. According to the intermodulation distortion analysis based on Volterra series, the reduction of the power amplifier linear gain is dependent on the feedback at the fundamental only, while the third-order intermodulation components are reduced due to effect of the feedback factor both at the fundamental and at the intermodulation frequencies when a reduction is equal to the loop gain in the latter case [52]. This means that, if the fundamentals were removed from the feedback loop, there would still be a reduction of the intermodulation products due to the feedback at the intermodulation frequencies with unaffected amplifier gain. In this case, the feedback signal is coupled from the output of the power amplifier, properly attenuated and phase shifted to form the error signal with cancelled fundamental by combining with a lower portion of undistorted input signal. The resulting error signal is then amplified, scaled in the amplitude and phase, and finally combined with a delayed upper portion of the undistorted signal to form the composite signal at the amplifier input with out-of-phased distortion products required to cancel the intermodulation products at the amplifier output. The stability analysis shows that, for narrow-band amplifiers with minimum loop delay, the reduction of third-order intermodulation products can exceed 20 dB [51].

The feedback linearization technique can also be implemented reversely to a feedforward scheme, as shown in Figure 16, where the linearizing circuit consists of the feedback, canceling and feeding blocks [53]. At the canceling block, the amplitude-corrected and phase-shifted feedback signal is combined with the sampled input signal to form the error signal, which is then amplified and added with proper amplitude and phase to the input undistorted signal within the feeding block. As a result, for a 38-dB main amplifier gain, the cancellation of the third-order intermodulation products from –22 to –42 dBc were achieved at the output power P_{1dB} = 27 dBm and operating frequency of 1.85 GHz. However, the critical problems of this analog feedback predistortion scheme are the bandwidth limitation caused by the loop delay and an oscillation tendency caused by the feedback nature. By employing a digital lookup table (LUT) technique, these limitations can be overcome, while maintaining the advantages of the feedback circuit [54]. As a result, the distortion is corrected in a digital domain and further enhanced by the feedback linearization. In this case, the structure of a digital feedback linearizer is the same as the analog feedback counterpart, except that the feedback signal in the cancellation loop constructs a LUT in the digital domain, and the gain factors of the signal canceling and feedback paths are adjusted by the DSP instead of using the variable attenuators and phase shifters. The predistortion signal is extracted directly from the LUT, which has been updated using the error signal extracted at the signal cancellation loop beforehand. Thus, the time delay through the loop is eliminated, and the bandwidth limitation does not exist anymore. At the same time, the oscillation tendency of the feedback

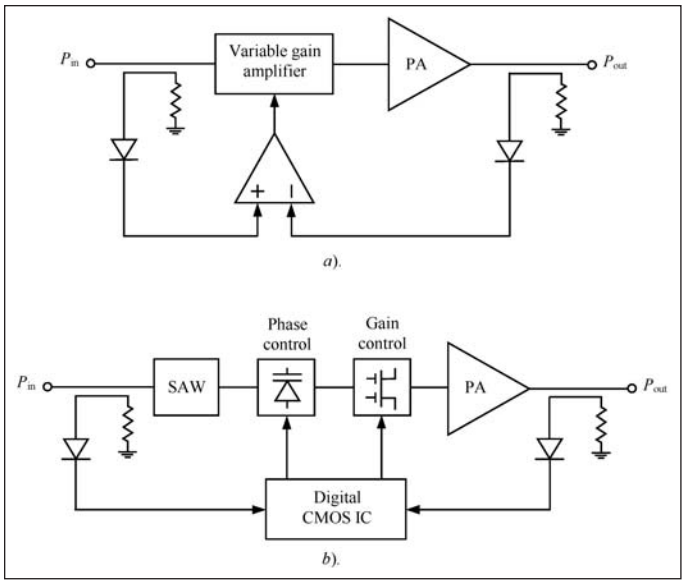

Figure 17 · Block diagrams of power amplifier with adaptive predistortion.

circuit can be suppressed easily by digital control of the feedback loop parameters. Compared to the conventional digital predistortion technique, less number of iterations is required for convergence with simpler algorithm.

The envelope-feedback linearization is a simple and popular technique to improve the distortion associated with the amplitude nonlinearity of the power amplifier [44, 55]. The basis of this technique, shown in Figure 17(a), is to compare the envelopes of the input and output signals, and to control the instantaneous gain of the power amplifier so as to minimize the difference between them. In this case, the RF input and output signals are sampled by the corresponding input and output couplers and then each fed to the proper input of the differential amplifier. The difference signal, representing the error between the input and output envelopes, is used to drive a variable gain amplifier to modify the envelope of the input signal which drives the power amplifier. The linearization effect depends on the nonlinearity of the detectors, especially at low signal levels, the bandwidth, time delay and phase-gain characteristic of the feedback loop and the sensitivity of the variable gain amplifier. Therefore, where the *AM-AM* distortion is dominant, the two-tone intermodulation products are typically reduced by up to 10 dB. To provide a significant improvement in the nonlinear distortion cancellation, both RF feedback and envelope feedback methods can be combined [44].

Figure 17(b) shows the envelope-feedback power amplifier module with a digital adaptive predistortion developed for CDMA handset application in a frequency range of 887-925 MHz [56]. The block diagram of the power amplifier module includes additionally a SAW filter, a phase-controlling block and a CMOS integrated circuit incorporating two look-up tables to linearize *AM-AM* and *AM-PM* characteristics. The variable gain amplifier is based on a dual-gate MOSFET to linearize *AM-AM* characteristic, which can easily vary the power gain by controlling its second-gate bias voltage in a range of more than 10 dB. In the phase-controlling block, a varactor diode was used with the phase range of more than 10°. By producing appropriate *AM-PM* predistortion data including phase characteristics of the dual-gate MOSFET and the following PA block, the total *AM-PM* can be linearized. Because the variations of phase with supply voltage and temperature according to the measurement results were insignificant, it was enough to use the lookup table with modified initial data to minimize phase variations versus input power that were predictable in advance. To linearize *AM-AM* characteristic, the adaptive predistortion was used by modifying the data in look-up table during linearization process. It was found that the allowable time delay must be less than 40 ns for CDMA signal. As a result, by using a digital adaptive predistortion mechanism for only *AM-AM* characteristic, the *PAE* was increased up to 48% for the output power of 27.5 dBm and *ACPR* of –49 dBc. The CMOS integrated circuit, whose size is 2.5×2.5 mm^2, consumed of about 15 mA.

The *AM-AM* linearization can also be achieved by dynamically varying the gate bias voltage of the final-stage transistor. In such an adaptive double-envelope feedback two-stage MESFET power amplifier the gain variations are detected directly while the phase variations are detected through a 90° branch-line coupler [57]. This is possible since, when the gate goes more negative, the MESFET is closer to pinch-off and its gain is reduced. On the other hand, as the gate goes more positive, the MESFET will approach Class A, where its gain is maximal. The dynamic bias on the gate of the device resulted in a 1 dB increase in P_{1dB} which improved the *PAE* by 5%.

To overcome the limitation of the envelope-feedback technique to correct for *AM-PM* distortion, the polar-loop technique can be used where a phased-lock loop is added to the envelope feedback system resulting in a polar-loop feedback [58]. However, the key disadvantage of a polar feedback occurs in the generally different bandwidths required for the amplitude and phase feedback paths, as well as locking capability of the phased-lock loop is limited at low signal level in the presence of interference coupled to the transmitter output from antenna that leads to a poorer overall performance. The Cartesian-feedback technique can solve at some extent the problems associated with the wide bandwidth of the signal phase by applying modulation feedback in the in-phase (I) and quadrature (Q) Cartesian baseband components. Figure 18(a) shows the basic block diagram of the Cartesian loop transmitter with two identical feedback loops for I and Q chan-

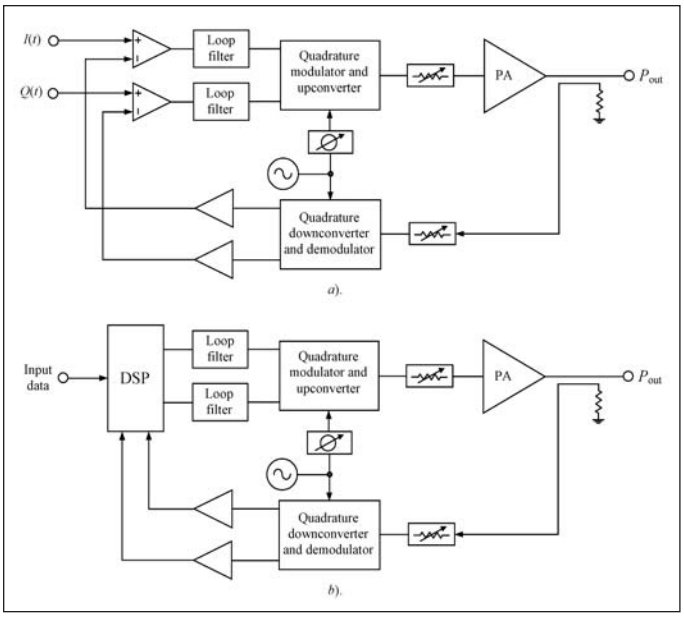

Figure 18 · Digital predistortion system.

nels [12, 59]. In this case, the sampled output signal is quadrature downconverted and demodulated to Cartesian co-ordinate signals, amplified and then subtracted from the corresponding input signals to form a distorted error signal which is complementary to the distortion signal at the PA output after being filtered and upconverted. The amount of the distortion reduction is equal to the loop gain and is high inside the passband of the loop filters. The phase shifter is necessary to control the phase delays which cause rotation of the signal constellation. The bandwidth of the Cartesian components is narrower than that of the RF signal, however errors in the feedback loop are not corrected and feedback components must be as linear as the desired loop linearity. The Cartesian feedback can be used to linearize multicarrier power amplifiers, improving output spectrum by 10-30 dB [60].

Further linearity improvement can be achieved by using a combined Cartesian loop and adaptive baseband predistortion linearization techniques using DSP, as shown in Fig. 18(b) [61]. Adaptive predistortion provides a continuous adjusting of the loop time delays and updating of the predistorter lookup table with information of the PA nonlinearity, which changes with supply voltage, temperature, load voltage standing wave ratio (*VSWR*), output power, and other environmental effects. For example, by using the predistorter implemented in DSP with 5 bit effective resolution digital-to-analog converters at its output and optimized using a direct search algorithm, a further 11 dB reduction of the third-order intermodulation components was achieved when a predistorted component was added to the error signal [62]. However, as the bandwidth of transmitting signal gets wider, the feedback loop delay mismatch becomes increasingly detrimental to the convergence of LUT adaptation algorithm. In this case, the analog Cartesian feedback, which is separated from LUT, can be used only for LUT training, thus resulting in the energy-efficient and low-complexity adaptive linearization [63]. Then, after the completion of LUT training, the analog Cartesian is turned off so that open-loop predistortion is performed using a compact Cartesian LUT to linearize wideband signals. The renewal of the LUT can happen without interrupting ongoing communication because the LUT training needs a millisecond and most communication protocols implement enough buffering for an error control method such as automatic repeat request. How often the LUT would need a renewal depends on PA characteristics and the environmental conditions.

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Author Information

Andrei Grebennikov received the MSc degree in electronics from Moscow Institute of Physics and Technology, and the Ph.D. degree in radio engineering from Moscow Technical University of Communications and Informatics. He can be reached by e-mail at: grandrei@ ieee.org

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Guidelines for Choosing RF/Microwave Coaxial Cable Assemblies

By Gary Breed
Editorial Director

Here is a review of the basic data used to specify cable assemblies for communications systems, test equipment and high performance, high reliability applications such as space

In the process of designing and constructing RF and microwave systems, few components are more common than cable assemblies. In most communication facilities, there are many cables used to interconnect the various pieces of equipment. Within those "boxes" cables will be used to route signals among the various boards and modules. And of course, the connection to the antenna system also requires cables.

In addition to system interconnections, the test equipment used during the design, manufacture, installation and commissioning of these systems use cables to deliver and receive their test signals. Many ongoing operational test and monitoring functions also send the acquired signal samples to the monitoring equipment via cables.

Whenever possible, it is best to use manufactured cable assemblies. They are built under controlled conditions, by experienced technicians, often using special fixtures for maximum precision and repeatability. Of course, many cables need to be measured on site, with connectors installed in the field. In these situations, the experience of the installer, along with the available tools and fixtures, will determine the performance of the final assembly.

Selecting the Cable and Connectors

Now that we have firmly established the widespread use of cable assemblies, a review of the basic information necessary to specify them is in order. The first step is to choose the right cable/connector combination to meet your needs. This is far more detailed than many new engineers expect. The following list is adapted from the user-input form on the Semflex Web site and shows the range of selection criteria for cable assemblies:

Application Data
- Cabinet interconnects (Y/N)
- Cable Type
  - (flexible, semi-rigid, hand-formable)
- Low Intermodulation (Y/N)
- Outdoor / Field Uses (Y/N)
- Chemical Resistance Conditions (Y/N)
- Cable Jackets and Outer Covers (Y/N)
- Cable diameter (0.120 to 0.500)
- Jackets:
  - (FEP, Polyurethane, Armor, Ruggedized, Nomex)

Electrical Specifications
- Maximum Frequency
- VSWR @ Max Freq
- Low Insertion Loss (Y/N)
- Impedance (ohms)
- Shielding Effectiveness @Max Freq.
- Power Handling
- Conductor Types
  - (Solid, stranded, silver plated)

Environmental Requirements
- Temperature Range
- Temp/Humidity (High/Med/Low)
- Vibration Shock (High/Med/Low)
- Abrasion (High/Med/Low)

Connectors
- SMA
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- TNC

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CBL-2FT-SMSM+	SMA	2	1.1	27	69.95
CBL-3FT-SMSM+	SMA	3	1.5	27	72.95
CBL-4FT-SMSM+	SMA	4	1.6	27	75.95
CBL-5FT-SMSM+	SMA	5	2.5	27	77.95
CBL-6FT-SMSM+	SMA	6	3.0	27	79.95
CBL-10FT-SMSM+	SMA	10	4.8	27	87.95
CBL-12FT-SMSM+	SMA	12	5.9	27	91.95
CBL-15FT-SMSM+	SMA	15	7.3	27	100.95
CBL-25FT-SMSM+	SMA	25	11.7	27	139.95
CBL-2FT-SMNM+	SMA to N-Type	2	1.1	27	99.95
CBL-3FT-SMNM+	SMA to N-Type	3	1.5	27	104.95
CBL-4FT-SMNM+	SMA to N-Type	4	1.6	27	112.95
CBL-6FT-SMNM+	SMA to N-Type	6	3.0	27	114.95
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CBL-2FT-NMNM+	N-Type	2	1.1	27	102.95
CBL-3FT-NMNM+	N-Type	3	1.5	27	105.95
CBL-6FT-NMNM+	N-Type	6	3.0	27	112.95
CBL-10FT-NMNM+	N-Type	10	4.7	27	156.95
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CBL-25FT-NMNM+	N-Type	25	11.7	27	199.95
			Female to Male		
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CBL-2FT-SFNM+	SMA-F to N-M	2	1.1	27	119.95
CBL-3FT-SFNM+	SMA-F to N-M	3	1.5	27	124.95
CBL-6FT-SFNM+	SMA-F to N-M	6	3.0	27	146.95
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Upcoming Web Classroom Courses:
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delivered via web classroom™

Summary:

This seminar discusses techniques for identifying the sources of unwanted coupling and radiation, and systematic approaches for their minimization.

Target Audience:

Product and package design engineers of all expertise levels will benefit from this course. A basic engineering background (BSEE or equivalent) is required.

Outline:

Electomagnetic Compatibility

• Definition of EMC • Ground definitions • Measuring Radiation : large scale (test box) and localized (probes)

Coupling and Radiation

• Differential and common mode connections • Coupling control techniques • Crosstalk between microstrip vs other lines • Radiation vs. loop area • Isolation techniques

Shielding

• Shielding effectiveness and wave impedance; shield materials • Effects of slots and holes in shield • Multiple small or fewer large holes? • Transfer impedance as the effectiveness parameter

Grounding

• Current distribution between shield and ground plane • Coupling through common ground inductance • Shared vias • Ground plane discontinuities and inductance effects • Choking off ground currents, filtering the power lines

Experimental/Numerical Techniques of Problem Solving & Case Studies

Monday	Tuesday	Wednesday	Thursday	Friday
Oct 26	Oct 27	Oct 28	Oct 29	Oct 30
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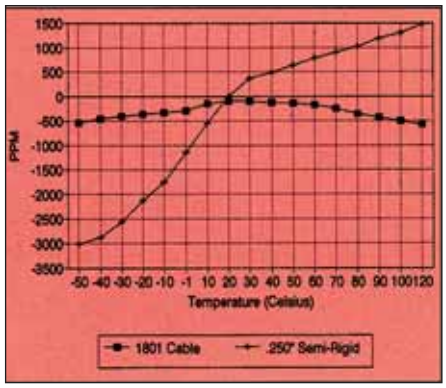


Figure 1 · Phase vs. temperature plot comparing standard 0.25 inch semi-rigid line with IW's 1801 phase stabilized cable.

Figure 2 · The Tuff-Grip adapters from Times Microwave Systems provide physical assistance for the attachment of field test cables.

High Performance Options

Beyond these basic selections, cable assemblies often have special performance requirements. Stable phase over temperature, exceptional flexibility, extreme temperatures, and repeated mating/unmating are a few of these.

For example, Storm Products provides phase stable cables according to the following customer choices (in increasing order of cost and manufacturing difficulty):

Relative phase match—a set of cable assemblies are matched to one another. Failure of one requires that all be replaced.

Absolute phase match—Cables are built to an exact standard. This is more expensive, but individual cables can be replaced.

Electrical length match over temperature—Should be used only for critical applications where the operating temperature cannot be controlled. Especially difficult to verify with short cables.

Phase stabilized cables have become more common as systems have increased in complexity, especially wireless base station equipment installed in unmanned locations. Phase stability over temperature assures performance over a wide enough range that precision air conditioning is not required. Figure 1 is a comparison by IW Microwave of standard semi-rigid cable with a phase stabilized cable.

Another high performance application is space-qualified operation. There are 50 or more applicable standards for space applications, developed by government, military, commercial and industry organizations.

Many of the necessary specifications are the same as any other cable assembly, but there are also many additional factors. According to W.L. Gore's extensive catalog of spaceflight cable assemblies, these include such things as outgassing in a vacuum environment, a temperature range of –55° to +150° C, ability to withstand launch shock and vibration, and a series of high-reliability practices including X-ray inspection.

Field-Built Assemblies

Wireless personal communications base stations make up the single largest consumer group for coaxial cable. While the base station equipment may use manufactured cable assemblies for its installation, the use of multiple antennas located at the top of towers requires that many cables are measured and cut on site, with field-installed connectors.

To address the need for maintaining high performance and precision installation in the field, cable and connector manufacturers offer appropriate tools and fixtures. These accessories allow the installation technicians to assemble connectors to cables with a precision and reliability that approaches in-house manufacturing.

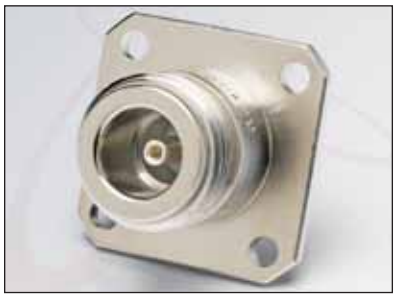
Test Cables

A special subset of cable assemblies are test cables, which are designed for repeated usage, high flexibility, and reliable connection over many mating/unmating cycles. Specifications are best determined by consulting the user's manual or applications support staff at the test equipment manufacturer.

Field testing requires yet another group of cables and accessories—special adapters. Broadcast, mobile radio and wireless base station equipment often uses large-size coaxial cables and connectors, while test equipment generally has Type N, SMA or other small-size connectors with well-defined characteristics to very high frequencies. Thus, adapters are an essential part of field testing. Among the products offered is a special grip from Times Microwave Systems, shown in Figure 2.

Summary

These basics should help you begin the process when specifying cable assemblies for your module, system or test application.



Type N Panel Receptacles

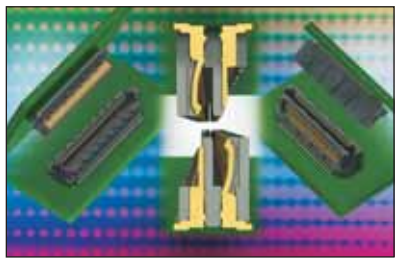
San-tron, Inc. has announced the release of a UG-58 A/U Type N panel receptacle that features single piece body construction. The seamless UG-58s are manufactured using a custom machine designed and built by San-tron for fully automated assembly. The UG-58 A/U panel receptacles have a frequency range of DC-11 GHz and a nominal impedance of 50 ohms. They are rated for operating voltages of 1000 V RMS and have a maximum dielectric withstanding voltage of 2500 V RMS at 60 Hz at sea level. The receptacles are designed for a temperature range of –65°C to +165°C and meet interface standards in accordance with MIL-STD-348. Pricing for UG-58 A/U panel receptacles start at $1.49 in quantities of 3,000+.

San-tron, Inc.
www.santron.com

SSPA and Transmitter

Introducing MITEQ's new medium power X-band rack-mount solid state power amplifier (SSPA) and transmitter systems. The new PA-R Model series is an SSPA rack-mounted system designed using a modular approach to provide flexible solutions to meet a variety of applications. These SSPA systems provide for over-temperature, overcurrent and high output VSWR safety protection. Status and control is achieved via the remote RS422/485 bus or Ethernet interface port. An event log is continuously updated with time-stamped records of significant events. MITEQ offers solid-state power amplifiers in C- and Ku-bands, as well as antenna mounted outdoor configurations.

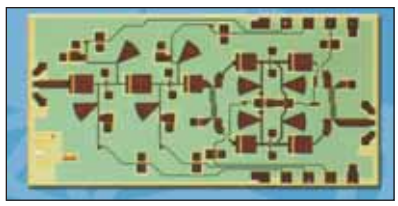
MITEQ, Inc.
www.miteq.com



Interconnect Systems

Samtec's high speed Razor Beam™ interconnect systems are designed to maximize PCB real estate with their slim, ultra-fine pitch, low profile designs. These space-saving systems are currently available in super-fine 0.4 mm pitch (ST4/SS4 Series) and self-mating 0.635 mm pitch (LSS Series) designs. A 0.5 mm pitch (LSHM Series) system for parallel, perpendicular, and coplanar applications is also in design. Razor Beam™ connectors are well suited for high speed applications with performance up to 10 GHz at –3 dB insertion loss. The systems are available in a choice of stack heights from 4 to 6 mm for the ST4/SS4 Series and from 6 to 12 mm for LSS Series and LSHM Series. Pricing begins at $ 0.054/per mated line.

Samtec, Inc.
www.samtec.com



GaAs HEMT LNA Die

Hittite Microwave Corporation has introduced a new GaAs HEMT MMIC low noise amplifier die, the HMC-ALH508, that will allow designers to address E-band applications from 71 to 86 GHz. The die is ideal for automotive radar, military and space applications, wireless LANs, and ultra high capacity E-band communication systems. The HMC-ALH508 is a three stage GaAs HEMT MMIC LNA chip that operates between 71 and 86 GHz, replacing the HMC-ALH459. It features 13 dB of small signal gain, 4.5 dB of noise figure and an output power of +7 dBm at 1 dB compression, and it operates from two supply voltages at 2.1 V and 2.4 V respectively, while consuming only 30 mA of supply current. HMC-ALH508 die samples are available from stock and can be ordered online.

Hittite Microwave Corporation
www.hittite.com

Flexible Coaxial Cable

Times Microwave Systems introduces the Phase Track II, an ultra-thermally stable flexible coaxial cable that can significantly improve performance in phase sensitive systems. The cable has the most phase stable dielectric material available to provide superior phase tracking performance from cable to cable with the lowest phase change through its temperature range. The phase performance achievement in Phase Track II cable is the result of a newly engineered dielectric material produced by Times Microwave Systems. The material called TF5 is thermally-mechanically balanced and is the most significant breakthrough in coaxial cable technology in decades. Phase Track II's TF5 is the most stable dielectric material available and virtually eliminates the wide phase changes that other high performance PTFE dielectric cables have with temperature changes.

Times Microwave Systems
www.timesmicrowave.com

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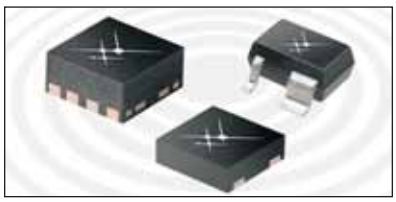
High Power Linear Amplifier

Aethercomm Model Number SSPA 0.4-0.6-100 is a high power, high efficiency, RF amplifier that operates from 400 to 600 MHz. This PA is ideal for military platforms as well as commercial applications because it is robust and offers high power over a 200 MHz bandwidth with excellent power added efficiency. This amplifier has a minimum P_{1dB} of 100 watts at room temperature. Noise figure at room temperature is 15.0 dB maximum. This amplifier offers a typical gain of 28 dB with a typical gain flatness of ±1.0 dB. Input and output VSWR is 2.0:1 maximum. Class AB quiescent current is ~1.5 amps typical employing a +28 Vdc supply. This PA operates from a +28 Vdc input voltage.

Aethercomm, Inc.
www.aethercomm.com



New Family of LNAs

Skyworks Solutions, Inc. introduced a new suite of low noise amplifiers (LNAs) featuring pHEMT and SiGe technologies. The new high performing, low cost product family serves various industry applications including GPS; infrastructure such as GSM, WCDMA and LTE base station transceivers; ISM band and satellite radio; and WLAN markets. The SKY65050-372LF transistor, the first released product in this family, is a high performance, low noise, n-channel, depletion mode pHEMT, fabricated from Skyworks' advanced pHEMT process and packaged in a miniature 4-lead SC-70 package. The SKY65050-372LF is available now and priced at 25 cents each in quantities of 10,000 devices.

Skyworks Solutions, Inc.
www.skyworksinc.com

Microwave Signal Generator

Agilent Technologies Inc. introduced a microwave signal generator with ultrahigh output power capability that breaks the 1-watt output barrier. The Agilent PSG E8257D signal generator option 521 delivers from +24 to +28 dBm specified output power over a frequency range of 250 MHz to 20 GHz (useable down to 10 MHz) and eliminates the need for additional hardware such as amplifiers, couplers and detectors. The Agilent PSG E8257D option 521 is priced at $41,575.

Agilent Technologies, Inc.
www.agilent.com

Filter Product Line

Achieving co-site interference protection in environments of multiple VHF/UHF tactical transceivers is now a reality through Pole/Zero's introduction of the Mega-Pole high power (50 watts) electronically tunable filter product line. When placed between the antenna and transceiver, the Mega-Pole yields excellent selectivity with minimal insertion loss, even in frequency hopping modes of operation.

Pole/Zero Corp.
www.polezero.com

Directional Couplers

MECA's 715 series directional couplers feature unique air-line construction that provides for the lowest possible insertion loss (less than 0.1 dB) high directivity (30 dB typ) and exceptional VSWR (1.10:1 typ) across the 0.800-2.200 GHz and 2.000-4.200 GHz bands. Rated for 500 watts average power (10 kW peak). Nominal coupling values of 6, 10, 20, 30 and 40 dB available from stock to 2 weeks.

MECA
www.e-MECA.com

Integrated Modulator

Analog Devices, Inc. (ADI) introduced a new highly integrated modulator for broadband satellite communications applications. ADI's new ADRF6750 modulator offers the highest level of integration and functionality by combining four devices into one small footprint. Operating within the 950 to 1575 MHz frequency range and on a single 5-volt power supply, the ADRF6750 modulator enables 10 Hz frequency increments, and the integrated output attenuator provides 47 dB output gain control with 1 dB step resolution. The device also features monitoring-output and output-disable-control capabilities. Furthermore, the ADRF6750 modulator offers best-in-class RF performance, including an output compression (P_{1dB}) of 8.5 dBm and output IP3 of 21 dBm. The ADRF6750 modulator is sampling now with volume production scheduled for July 2009. It is available for $5.75 per unit in 1000-unit quantities.

Analog Devices, Inc.
www.analog.com

Surface Mount Pin Diode Switches

Aeroflex/Metelics is pleased to announce the release of a family of surface mount, silicon PIN diode SP2T switches with power handling capability up to 100 W C.W. incident RF power. They are optimized to function in two frequency bands of interest: 20-700 MHz, and 400-4000 MHz, and are available in compact surface mount packages measuring 8 × 5 × 2.5 mm. The MSW2000 series of surface mount PIN diode SP2T switches is manufactured using Aeroflex / Metelics' proven hybrid manufacturing process incorporating high voltage PIN diodes and passive devices within a ceramic substrate. Both asymmetrical and symmetrical SP2T topologies are offered. These products are available in high volume tape-reel and tube through Aeroflex/Metelics distribution networks.

Aeroflex / Metelics
www.aeroflex.com/metelics

Fixed Frequency Synthesizer

Z-Communications, Inc. is pleased to announce a RoHS compliant fixed frequency synthesizer in X-band. The SFS11000Z-LF is a single frequency signal source that is pre-programmed to operate at 11,000 MHz and features a self locking mechanism with lock detect. This synthesizer has a typical phase noise of –84 dBc/Hz at 10 kHz offset with a typical reference spur suppression of –70 dBc. The SFS11000Z-LF is designed to deliver a typical output power of 0 dBm with a VCO voltage supply of 5 Vdc while drawing 90 mA (typical) and a PLL voltage of 3 Vdc while drawing 11 mA (typical) over the temperature range of –30° to 60° C. SFS11000Z-LF is well suited for broadband transmission applications that require low phase noise performance.

Z-Communications, Inc.
www.zcomm.com

Low VSWR Multiway Coax Connectors

Sabritec announces the launch of their new high frequency Micro-D Coax Connectors (MDHC) with low VSWR up to 40 GHz (typ). These connectors have a constant 50 ohm airline impedance interface and are 30% smaller than their SCX coaxial connectors. The ultra-miniauture MDHC contacts are compact in size allowing multiple high density connections in a Micro-D style shell for applications where tight space and weight savings are critical. MDHC Connectors are also available with locking post mechanisms or blind-mate hardware attachment. Insert arrangements are available in 2, 4, 6, 8 and 10 way coax assemblies as well as mixed signal, power, coax, and high speed twinax for hybrid arrangements.

Sabritec
www.sabritec.com

New Antenna Design Product

Computer Simulation Technology AG (CST), EM Software & Systems-S.A. (Pty) Ltd. announce the release of a new software product to accelerate the antenna design and modelling process. Antenna Magus has a huge database of antennas that can be explored to find, design and export models of designed antennas to CST MICROWAVE STUDIO® and FEKO. Its feature set is targeted at aiding engineers get to the

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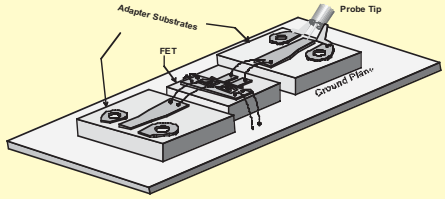

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Computer Simulation Technology
www.cst.com

Precode/Decode Software

Anritsu Company extends the 100 Gbps test capabilities of its industry leading MP1800A signal quality analyzer with the introduction of precode/decode software that has been developed to support the latest optical phase modulation schemes, including DP-QPSK, DQPSK, DPSK, and ODB, used in next generation networks. The precode and decode functions of the MX180000A-001/002 option help reduce 100G and 40G core-network R&D costs as well as time to market by supporting fully automatic generation of modulation signals needed to evaluate 100G DP-QPSK and 40G DQPSK, DPSK, and ODB optical modulation technologies. The precode/decode software package is priced at $1,500, while the base MP1800A for DQPSK is $257,000. Delivery is 6 to 8 weeks ARO.

Anritsu Company
www.us.anritsu.com

3D Planar EM Simulator

Sonnet Software, Inc. introduces Sonnet Lite Release 12, a free 3D Planar EM simulator, currently available for download, as a tool for learning and experimenting in high frequency RF transmission lines, circuits, packages and antennas. Release 12 is Sonnet's sixth full release of Sonnet Lite over the past decade and is a fully functional EM software tool that does not expire and is based on a proven industry standard in the Sonnet Professional Suite and introduces higher simulation efficiency and more features. New features in Release 12 will help high frequency design engineers explore manufacturing tolerance impacts on designs. Sonnet Lite can also be used from within the AWR Microwave Office® or Agilent ADS™ environments.

Sonnet Software
www.sonnetsoftware.com

Programmable Attenuator

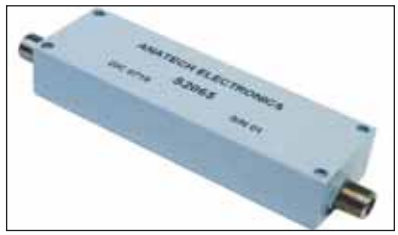
BroadWave Technologies has developed an electromechanical 75-ohm programmable attenuator with an external TTL control board and BNC female connectors. Model 674043063 offers a dynamic range of 63 dB in 1 dB steps over a frequency range of DC-2 GHz. Attenuation accuracy is ±0.5 dB or 2% of programmed, 1.50:1 maximum VSWR and 3.0 dB insertion loss. The switching speed is 10 milliseconds and TTL control comes via a 10-pin connector with a latch. Other impedance and connector options are also available.

BroadWave Technologies, Inc.
www.broadwavetech.com

Non-Linear Transistor Library

Modelithics, Inc. has announced the release of an enhanced version of the Modelithics NLT Library™ version 4.0, now with support for new TriQuint Semiconductor GaAs pHEMT and HFET models, among other updates. The Modelithics' NLT Library enables RF designers to model key performance parameters with high accuracy eliminating the expense or time necessary with PC board prototyping. The NLT 4.0 upgrade will be forwarded, free of charge, to all customers currently under a Modelithics Platinum Maintenance contract for either the NLT Library or the Modelithics Library Complete for ADS.

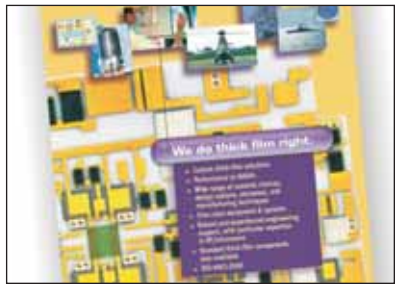
Modelithics, Inc.
www.modelithics.com



Notch Filter for TV Broadcast Applications

Anatech Electronics introduced the Model S2065 notch filter for use in the television broadcast band that reduces the level of undesired signal by providing an extremely high level of signal rejection. The Model S2065 has 3-dB passbands of DC to 750 MHz and 850 to 1200 MHz, rejection at 800 to 806 MHz greater than 70 dB, return loss greater than 14 dB. It handles an RF input power of 5 W and will operate over a temperature range of –40° C to +80° C. The Model S2065 is a custom product and can be specified with different electrical and physical characteristics.

Anatech Electronics, Inc.
www.anatechelectronics.com



New Thick-Film Literature

Anaren, Inc. announced that it has developed a free informational kit developed to introduce design engineers, manufacturing engineers, and purchasing professionals to the company's thick film ceramic circuits capabilities. Anaren's thick film service is offered through the company's Anaren Ceramics subsidiary, located in Salem, New Hampshire. The kit includes a design guide and information on the company's experience with thick-film across the defense, commercial wireless, and consumer electronics sectors.

Anaren, Inc.
www.anaren.com

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DESIGN NOTES

Feedback and Comments on "Where Have the Spaces Gone?"

We received many reader comments about last issue's column on the proper format for numbers and units. Here are excerpts from some of your e-mails, along with follow-up notes of our own.

Editor,

Thank you for your design note on spaces. I have fallen victim to leaving out the spaces. But I checked the official SI standard once again, and I find that paragraph 5.3.3 specifies that spaces are to be used:

http://www.bipm.org/utils/common/pdf/si_brochure_8_en.pdf

Thank you for the fine magazine.

Steven Swift
Novatech Instruments, Inc.

Editor,

Ref. High Frequency Electronics Design Notes, May 2009. Two online manuals I find quite helpful in technical writing are:

U.S. Government Printing Office Style Manual
http://www.gpoaccess.gov/stylemanual/browse.html

IEEE Standards Style Manual
http://standards.ieee.org/guides/style/

Dave Bowker
Fort Kent, ME

Editor,

Your Editorial/Design Notes on "Where Have the Spaces Gone?" resonated with me as an aesthete when it comes to the written word. Your specific question of when did the practice of removing spaces between numbers and units become common practice speculated that there was some reason for the change, but you could not find it. Let me offer a possibility based on my own observations. With the shift to self editing of documents (e.g., conference papers, web pages, etc.), those that do the writing do not know what the rules are, and since there is no editor to make the changes necessary, mistakes start to propagate.

I have noticed this with students when I review their papers and theses. When asked why they make these mistakes, they say they looked at previous students' work or conference papers. Unfortunately, they look at the wrong examples. So, I take it upon myself to get out the red pen and bleed all over their drafts until they get it right. They remember this and get it right from then on.

Sven G. Bilén, Ph.D., P.E.
Chief Technologist, Center for Space Research Programs
The Pennsylvania State University

Editor,

I suspect the missing spaces between numerals and units are caused by two things:

1. Writers don't know how to insert a "non-breaking space" in their word processors. Use of a regular space can result in numerals and units on different lines of text (another writing no-no that is frequently ignored, even in prestigious refereed journals).

2. With no space, there is no need to hyphenate adjectival uses (e.g., "5-V source" [correct] as opposed to "5 V source" [incorrect]).

I grade dozens of project reports from undergraduate EE students every semester. Whenever I think I've seen every possible grammatical and style error, one of my students invents a new one. It is far, far easier to ignore the rules of grammar and style than to try to teach them.

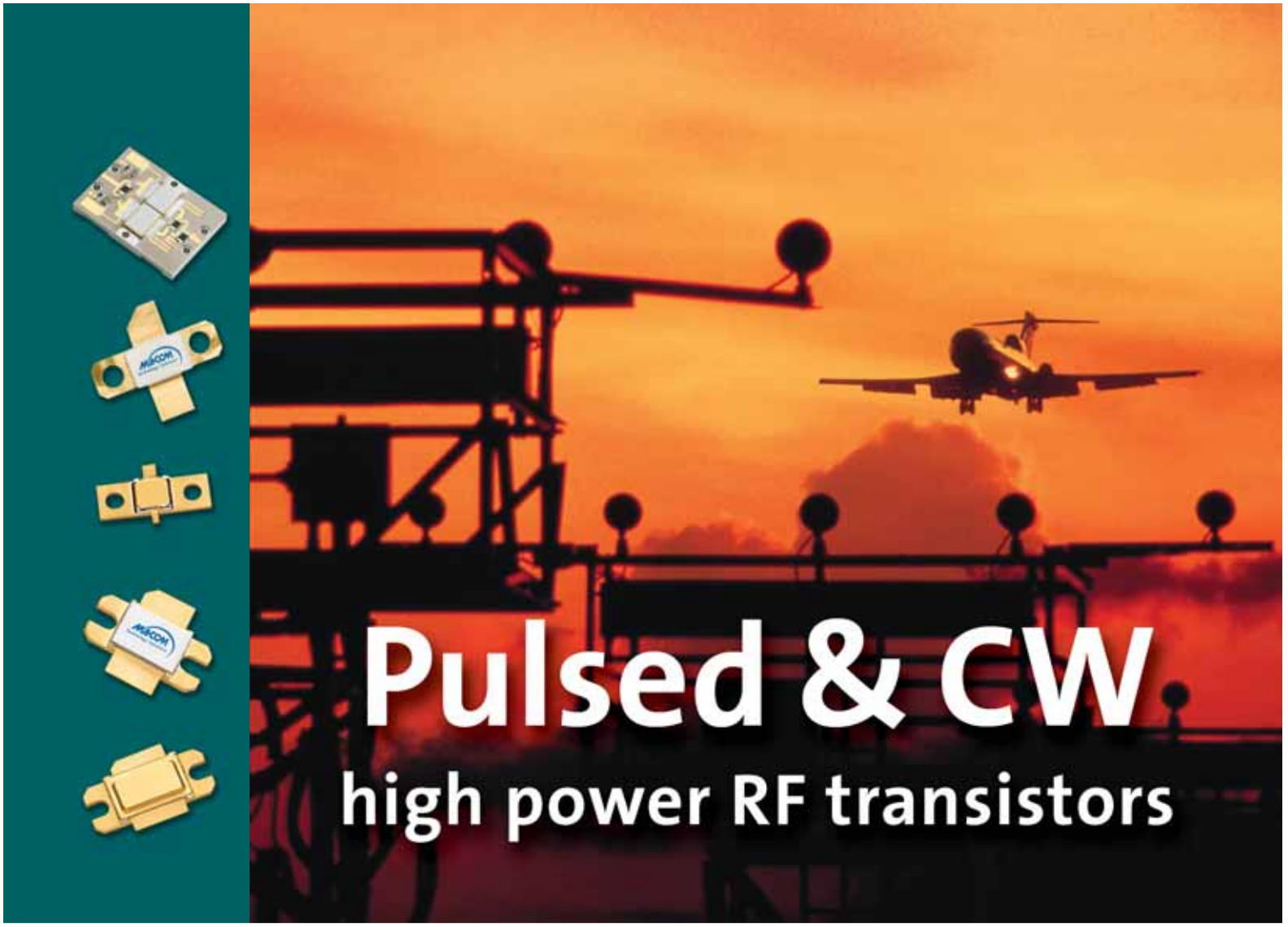
Prof. David M. Drury
University of Wisconsin–Platteville

Editor's Comments

David, thanks for mentioning the nonbreaking space, which is the key to keeping numbers and units together as the text wraps to the next line. In MS Word, use the command CTRL-SHIFT-SPACE, or look up the proper action in other programs.

I also apologize to grammatical experts everywhere for all the little errors we make while preparing a wide variety of material for each issue. For example, I know we miss a lot of those pesky hyphenated modifiers in our magazine! While "5-V source" is grammatically correct, it is very easy—even for professional editors—to look past them when the context is perfectly clear.

Finally, we still have not found any style guide that recommends the no-space usage. Any help to identify the origin of this practice and satisfy our curiosity would be appreciated! *—Gary Breed*



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PH2931-135S	2700-3100	135	7.5	40
*MAPP-002729-300M00	2700-2900	300	8.3	40
*PHA2729-240M	2700-2900	190	7.4	39
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MRF151G	2-175	300	14	50
MRF392	30-500	125	8	50
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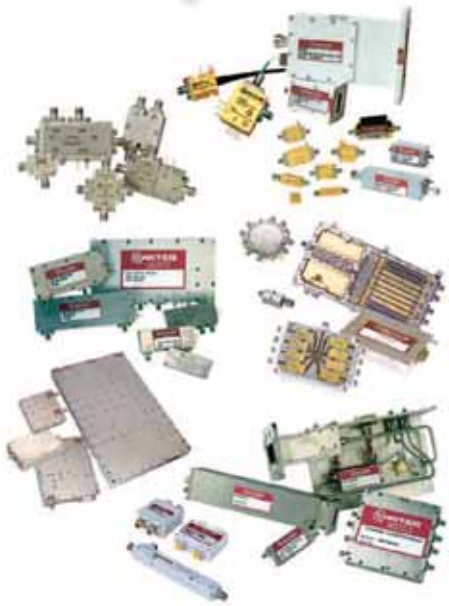

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